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An Implantable Transmitter and a Complete Telemetry System for Body Temperature Measurement

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AN IMPLANTABLE TRANSMITTER AND A

COMPLETE TELEMETRY SYSTEM FOR

BODY TEMPERATURE MEASUREMENT

BY

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COMPLETE TELEMETRY SYSTEM FOR

BODY TEMPERATURE MEASUREMENT

This thesis is approved as a creditable and independent investigation by a candidate for the degree, Master of Science, and is acceptable as meeting the thesis requirements for this degree. Acceptance of this thesis does not imply that the conclusions reached by the candidate are necessarily the conclusions of the major department.

Thesis Advisor Date

Acting Head, Electrical Engineering Department Date

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D.D.D.

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CHAPTER I

INTRODUCTION

This manuscript includes the specification, design, operation and results of an implantable temperature sensing transmitter and a pulse frequency to voltage converter with a digital temperature readout. The research project also includes Frequency Modulated (FM) receiver to complete the temperature measuring telemetry system.

1-1 THE PROBLEM

What are telemetry and biotelemetry? According to the Modern Dictionary of Electronics, telemetry is "the science of sensing and monitoring of information at some remote location and transmitting the data to a convenient location to be read or recorded".[1]

Biotelemetry is "the technique of measuring or monitoring vital processes and transmitting data without wire to a point remote from the subject".[1] Biotelemetry is also the formal name for radio tracking of wildlife.[2]

Biotransmitters are the transmitters used to measure, monitor and transmit vital or physiological processes such as electrocardiogram, electroencephalogram and body temperatures. Some biotransmitters are capable of transmitting all three of the above vital processes simultaneously but they are usually bulky, unreliable, and inaccurate. Therefore, at the present state of the art, it is desirable to design biotransmitters to measure one specific physiological parameter.

Why use biotelemetry? Biotelemetry involving wildlife has helped scientists to gain knowledge and thereby save some endangered species, slow the spread of animal disease and plan better food-and-cover areas for wildlife.[3]

Wildlife telemetry is getting so sophisticated that experts can detect the erratic flight of a sick bird. In Montana scientists have learned to recognize the spasms of a "bugged" blue grouse suddenly poisoned by insecticides.

Medical men are interested in the results of data collected from the animal tests because these tests may provide answers to giving man a healthier and better life. A quote from the Minneapolis Tribune Picture Magazine, Sunday, January 21, 1973, p.15 gives another reason for the use of biotelemetry:

> "Biotelemetry can be of enormous aid to us in understanding natures' restrictions", said Testes. "We must learn what the baseline is - what is normal, what are the boundaries of natures rules.
>
> Perhaps we can even gain some clues about how we should regulate our own population and society."

Why study body temperature with biotelemetry? With current day stresses on our environment, it has become necessary for the scientists to study the relation of the cause-effect of our pollutants, drugs, radiation and environment on the bodies of human beings and animals. Body temperature is an important physiological parameter in the study of the physical state of the body.

Scientists use temperature studies as a method of detecting certain effects on the body. For instance, body temperature studies can be used to detect such things as body activity, health, disease, psychological and physiological stresses, and drug effects.[4]

Scientists find it desirable to constantly monitor the subject's body temperature without restraints to minimize the subject's movement. This assures more accurate testing procedures since the subject may fight the restraints and thus generate body heat, generating erroneous body temperature measurements.

Wildlife scientists also are interested in monitoring animal body temperatures with the wildlife in their natural habitat. Biotelemetry would enable the study of the environmental influences on the body temperature of wildlife. Due to the present interest in temperature biotelemetry, a proposal for a complete temperature sensing telemetry system can be seen in the following section.

1-2 A PROPOSAL OF A TEMPERATURE MEASURING TELEMETRY SYSTEM

The temperature sensitive biotransmitter discussed in this manuscript shall be of the implantable type. To be implantable the biotransmitter must be encapsulated in a tissue compatible and fluid resistant material.

The temperature sensing unit (TSU) is a programmable unijunction circuit. The circuit has stability, simplicity, and extremely low current drain. A thermistor is the temperature sensing element of the circuit. Circuit linearity is essential between 32 and 40° centigrade. The circuit must have $\pm .05$ °C accuracy.

The transmitter frequency is to be near 150 MHz out of the commercial radio band. The TSU pulse modulates the frequency of the transmitter. FM is used because of a lower signal to noise ratio.[15]

The implanted transmission range for the biotransmitter is at least 100 feet.

A commercial FM receiver is used to monitor the biotransmitter. The sensitivity is approximately ± 0.25 microvolts. A more detailed discussion of the receiver is given in Chapter III.

A converting or decoding system is needed to change the pulse data to the corresponding temperature readings. The pulse frequency of the biotransmitter is proportional to the body temperature. Therefore the information available from the radio is a frequency modulated pulse train which corresponds to the subjects body temperature. A converter uses the pulses to trigger a clamped monostable multivibrator which produces an output pulse of fixed amplitude and width for each input pulse. These pulses are averaged in a low pass filter circuit. The dc voltage level is proportional to the frequency of the pulses of the biotransmitter or to the subjects' body temperature.

The temperature versus frequency plot of the TSU does not have a slope of one. In general each TSU will have a different slope. To have a direct frequency to temperature conversion, the frequency to temperature slope of the converter must be the same as that of the TSU. It is therefore, essential that the frequency versus temperature slope and the offset of the converter be adjustable. The design and description of the converter circuit are given in Chapter IV.

A digital voltmeter is used to monitor the dc voltage level and thus give a direct readout of temperature. The results of the digital voltmeter used in this research project are also included in Chapter IV.

CHAPTER II

TEMPERATURE SENSING UNIT

The requirements for a reliable TSU are as follows: Low power consumption, minimum number of components, small size, light weight, high sensitivity, good temperature stability, a stable power supply, and long life. The TSU circuit chosen for this research project was a relaxation type oscillator using a programmable unijunction transistor (PUT) in conjunction with a thermistor for temperature sensing. The circuit consists of four passive elements in addition to the PUT and the thermistor which make possible a compact and lightweight TSU construction. This type of circuit offers a current drain in the low microamperes, however the sensitivity depends on the type of thermistor used.

The PUT is a relatively new device. It is normally used in conventional unijunction transistors (UJT) circuits. Basically, the characteristics of both devices are similar, but the triggering voltage of the PUT is adjustable or programmable. The programming is accomplished by setting an external resistive voltage divider network. The PUT is more sensitive and faster than the UJT.[5] In general the PUT is a more versatile and economical device for most applications than the UJT.

2-1 UNIJUNCTION TRANSISTOR

To understand the operation of the PUT, it is helpful to know the theory of the UJT. The UJT (Fig. 2-1) has a small rod of P material (P-type crystals are doped with acceptor atoms) extending into the block of N material (N-type crystals are doped with donor atoms) which serves as a P-N junction.[6]

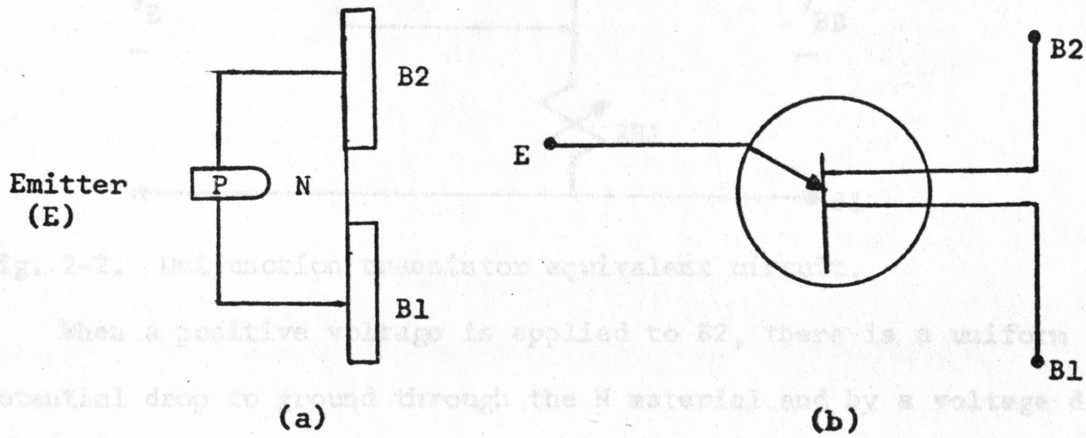


Fig. 2-1. Unijunction transistor: (a) construction; (b) symbol for unijunction transistor with N-type base.

Normally the B1 and B2 are two metallic contacts welded to the N block which create new junctions. The electrode base-one (B1) is the common return for the circuit. The interbase resistance (Rbb) of the N material between B1 and B2 is measured at room temperature by applying a positive voltage to B2 with the emitter open circuited. Rbb is usually between 4.7 Kilohms (K) and 9.1 K and is defined from Fig. 2-2 as follows:[7,8]

$$Rbb = RB1 + RB2. \qquad (2\text{-}1)$$

Rbb is temperature dependent and it increases linearly with temperature up to about 140° centigrade. The temperature coefficient at 25° centigrade is about + .08%/degree centigrade.[9]

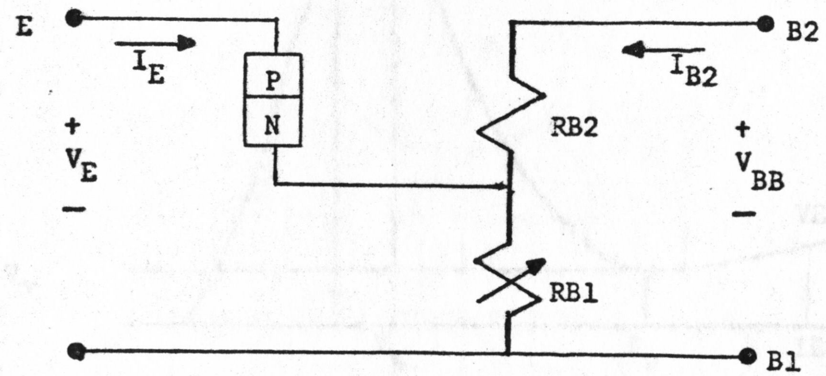


Fig. 2-2. Unijunction transistor equivalent circuit.

When a positive voltage is applied to B2, there is a uniform potential drop to ground through the N material and by a voltage divider action, the open emitter voltage V_E is theoretically $\eta\ V_{BB}$. The coefficient η is known as the intrinsic standoff ratio and the voltage divider is given by the following equation:

$$\eta = \frac{RB1}{RB1 + RB2} . \qquad (2\text{-}2)$$

Typical values for η are between .50 and .80.

When the diode (PN junction) of the unijunction transistor becomes forward biased, the RB1 resistance changes to a lower resistance value. This generates the negative resistance characteristic between emitter E and B1 (Fig. 2-3). As shown in Fig. 2-3, very little emitter current flows until the emitter voltage V_E is raised sufficiently to forward bias the diode. The peak point voltage V_P (Fig. 2-3) is the emitter voltage at which the diode becomes forward biased. The peak point voltage is equal to a fraction of V_{BB} plus the diode drop V_D.

$$V_P = \eta V_{BB} + V_D \qquad (2\text{-}3)$$

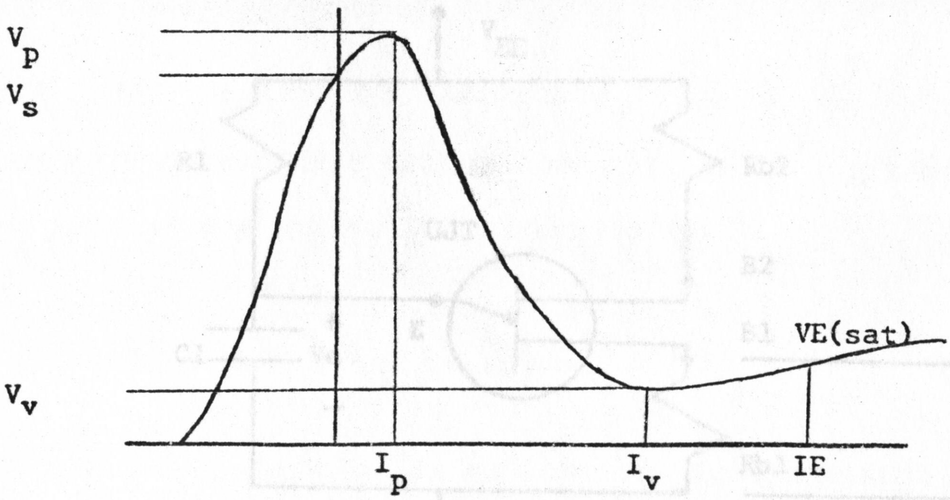


Fig. 2-3. Unijunction transistor static negative resistance characteristics.

The voltage V_D is about 0.7 volt at 25° centigrade and decreases about 2 milivolt/degree centigrade.[9] At the voltage V_P, the diode switches or fires. With higher emitter currents, the emitter to B1 voltage drops, increases, and levels off to saturation value $V_{E(sat)}$.

Armed now with a basic understanding of the UJT operation, we are ready to discuss the operation of the UJT relaxation oscillator. The UJT and the PUT relaxation oscillator are basically the same; except several resistors normally needed to bypass inter-base current and for temperature compensation on the UJT circuit may be left out of the PUT circuit.

Initially consider the UJT to be in the off state in the UJT relaxation oscillator (Fig. 2-4). In order for the oscillator circuit to function, the load line drawn between V_{EE} and $V_{EE}/R1$ must intersect the characteristic negative resistive region which is between the V_p voltage and valley V_v voltage as shown in Fig. 2-5.

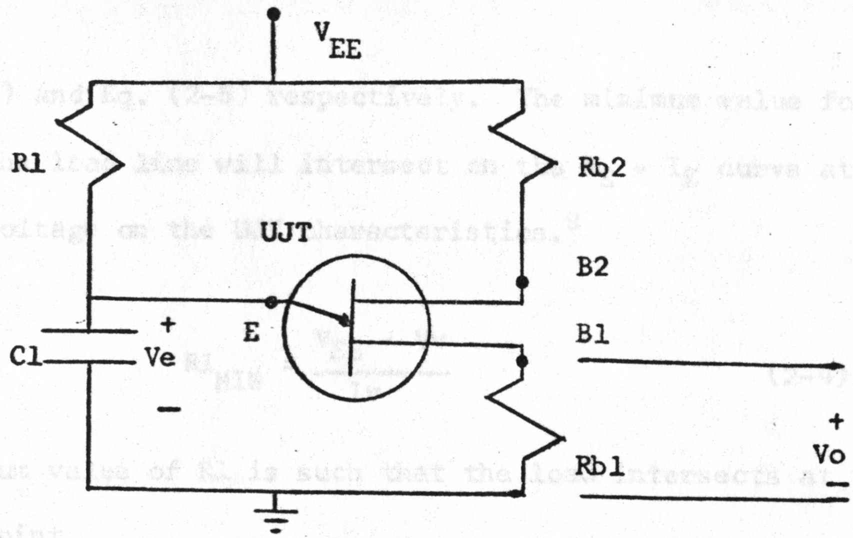


Fig. 2-4. Unijunction transistor relaxation oscillator.

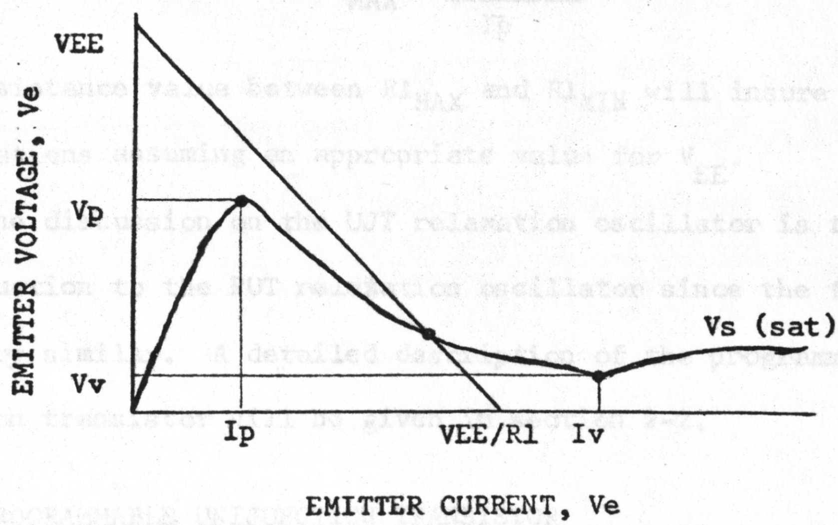


Fig. 2-5. Load line for the static characteristics of a unijunction relaxation oscillator.

The capacitor C1 charges through resistor R1 until a voltage greater than Vp is reached. The transistor will then turn on, discharging C1 through B1 and developing an output impulse voltage across resistor Rb1.[6] The high emitter current is limited by Rb1 and a small emitter impedance.

R1 is a control resistor for determining the load line for an astable situation. Minimum and maximum values for R1 can be calculated using Eq. (2-4) and Eq. (2-5) respectively. The minimum value for R1 is such that the load line will intersect on the V_E - I_E curve at the valley point voltage on the UJT characteristics.[9]

$$R1_{MIN} = \frac{V_{EE} - Vv}{Iv} \qquad (2\text{-}4)$$

The maximum value of R1 is such that the load intersects at the peak voltage point.

$$R1_{MAX} = \frac{V_{EE} - Vp}{Ip} \qquad (2\text{-}5)$$

Any resistance value between $R1_{MAX}$ and $R1_{MIN}$ will insure astable oscillations assuming an appropriate value for V_{EE}.

The discussion on the UJT relaxation oscillator is to serve as an introduction to the PUT relaxation oscillator since the two circuits are very similar. A detailed description of the programmable unijunction transistor will be given in section 2-2.

2-2 PROGRAMMABLE UNIJUNCTION TRANSISTOR

The programmable unijunction is a three terminal passivated PNPN device (Fig. 2-6a).[7] The terminals are designated as ANODE, ANODE GATE and CATHODE.

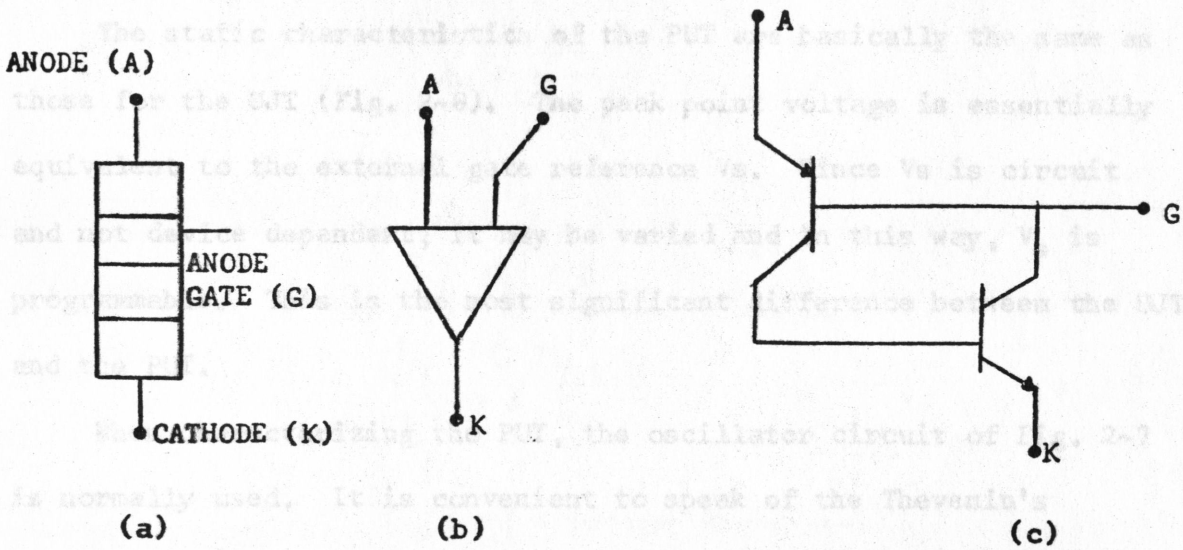


Fig. 2-6. Programmable unijunction transistor. (a) Construction (b) Symbol (c) Transistor equivalent

A typical PUT relaxation oscillator circuit is shown in Fig. 2-7. To operate, the PUT must have a gate reference voltage (Vs). A voltage divider is usually used to bias the voltage Vs.

$$Vs = \frac{VccR1}{R1 + R2} \qquad (2\text{-}6)$$

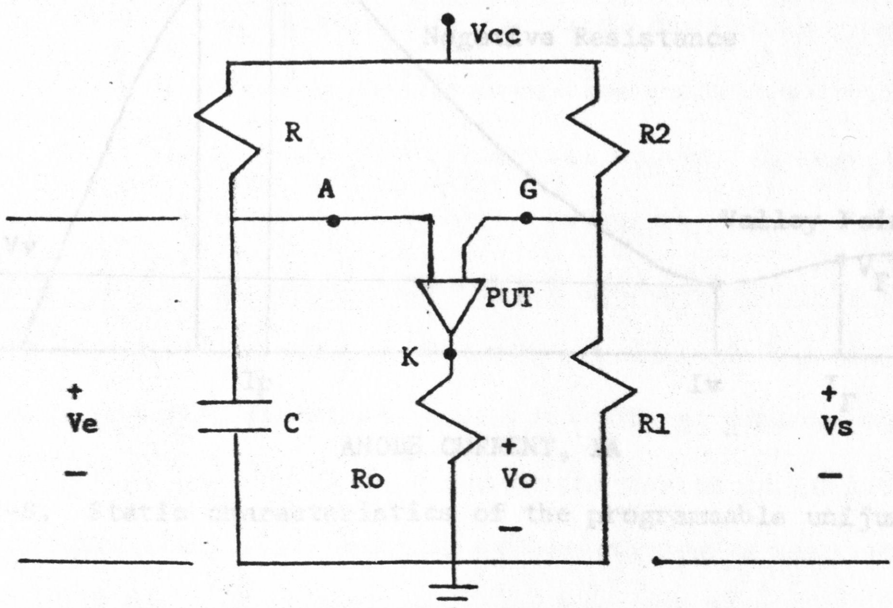


Fig. 2-7. Typical programmable unijunction transistor oscillator circuit.

The static characteristics of the PUT are basically the same as those for the UJT (Fig. 2-8). The peak point voltage is essentially equivalent to the external gate reference Vs. Since Vs is circuit and not device dependent, it may be varied and in this way, V_s is programmable. This is the most significant difference between the UJT and the PUT.

When characterizing the PUT, the oscillator circuit of Fig. 2-7 is normally used. It is convenient to speak of the Thevenin's equivalent circuit for the external gate voltage Vs and the equivalent gate resistance Rg. Vs has been defined in Eq. (2-6) and Rg is given by Eq. (2-7). The equivalent circuit is shown in Fig. 2-9.

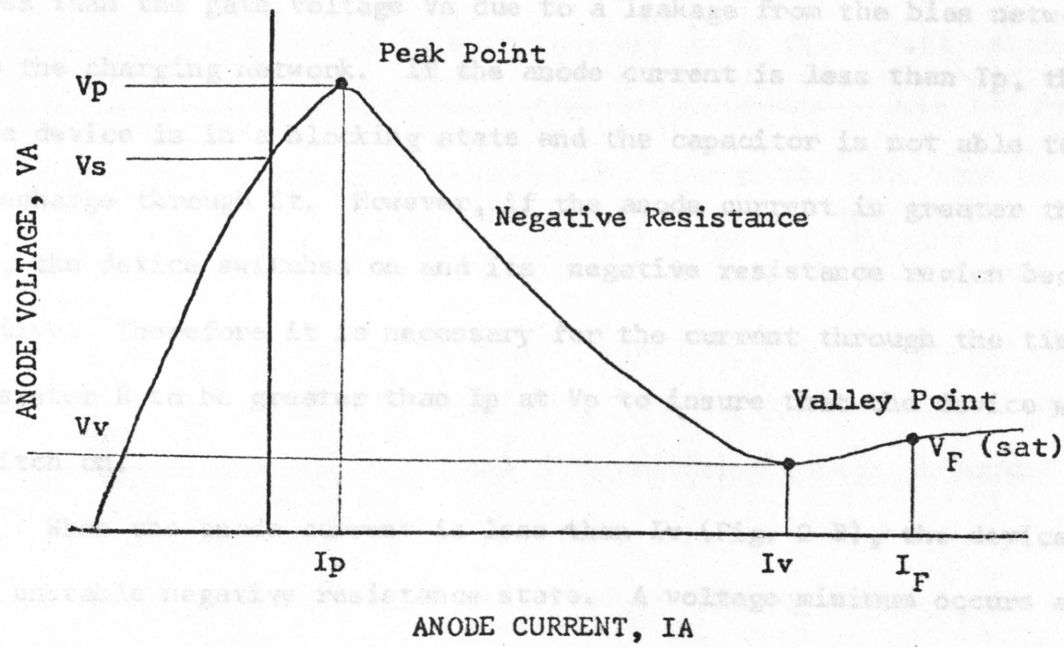


Fig. 2-8. Static characteristics of the programmable unijunction.

$$Rg = \frac{R1R2}{R1 + R2} \tag{2-7}$$

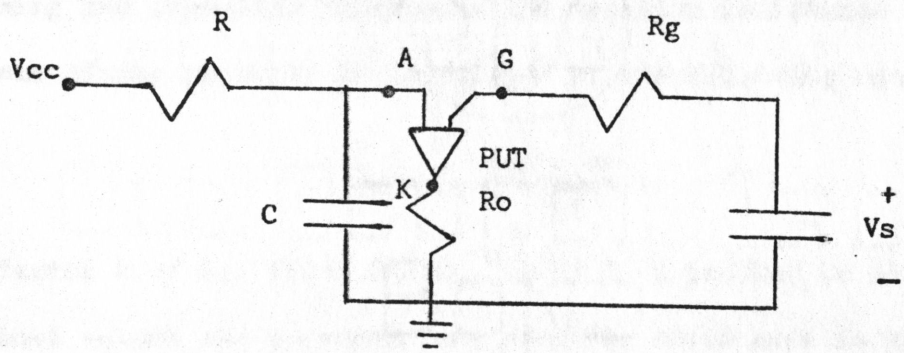


Fig. 2-9. Equivalent circuit for the PUT relaxation oscillator.

The peak point is shown graphically by the static characteristic curve of Fig. 2-8. A reverse anode current flows with the anode voltage less than the gate voltage Vs due to a leakage from the bias network to the charging network. If the anode current is less than Ip, then the device is in a blocking state and the capacitor is not able to discharge through it. However, if the anode current is greater than Ip, the device switches on and its negative resistance region becomes active. Therefore it is necessary for the current through the timing resistor R to be greater than Ip at Vp to insure that the device will switch on.

When the anode current is less than Iv (Fig. 2-8), the device is an unstable negative resistance state. A voltage minimum occurs at Iv. An anode current greater than Iv causes the device to go into a "on" state or saturation. The charging current through R must be less than Iv at the valley point voltage to guarantee oscillation.

Armed with these limits we can determine the range of values for the timing resistor. The minimum value for R is

$$R_{MIN} \geq \frac{2(Vcc - Vv)}{Iv} \qquad (2\text{-}8)$$

assuming the load-line intersects the negative resistance region. The maximum timing resistor is calculated by the following equation:

$$R_{MAX} \leq \frac{Vcc - Vp}{2Ip} . \qquad (2\text{-}9)$$

The factor 2 of Eq. (2-8) and Eq. (2-9) is a padding to avoid the critical values and to absorb any possible resistance in the capacitor.

As mentioned earlier, the peak point voltage can be determined externally. The expression used to describe peak point voltage is

$$Vp \quad V_T + V_s, \qquad (2\text{-}10)$$

where Vs is the external gate voltage and V_T is the offset voltage.[5] The offset voltage is the combination of the anode-to-gate (diode) voltage and voltage caused by the flow of Ip at the gate just prior to triggering. Therefore the equation for the offset voltage is

$$V_T = V_{AG} + I_p R_g . \qquad (2\text{-}11)$$

Normally V_T is given in the electrical characteristics as a function of V_p and dynamic impedance, but also switching speed. If small capacitors (less than .01 µF) are used, then part of the charge is lost during the turn on interval. The use of relatively large capacitors (approx. 0.2 µF) will reduce this type of loss and will increase V_o.

Just as in the basic circuit operation of the unijunction oscillator of section 2-1, the capacitor charges through R until the voltage Ve (Fig. 2-10a, pts.A-B) is greater than Vp. An equation for the charging process can be given as follows:[10]

$$Ve_{A-B} = Vcc(1 - e^{-(t_1/RC)}), \qquad (2\text{-}12)$$

$$e = 2.71828.$$

The capacitor will discharge through the anode after Ve is greater than Vp. The anode voltage waveform would correspond to the B-C portion in Fig. 2-10a. At this time the PUT becomes a low impedance (approx. 10 ohms) so the current limiter is Ro.[9] The B-C portion of Fig. 2-10a can be described by the following equation:[10]

$$Ve_{B-C} = Vpe(t_2-t_1)/\tau_1, \qquad (2\text{-}13)$$

$$\tau_1 = (R1 + 10)C$$

The output voltage Vo waveform (Fig. 2-10b) is very similar to the t_1 - t_2 portion in Fig. 2-10a because Eq. 2-13 also describes the t_1 - t_2 portion of Vo. In order for Eq. 2-13 to apply for Vo, Vp must be substituted by Vo_{MAX}.

The programmable unijunction transistors that were chosen for this research project were the GE D13T2 and an equivalent Motorola 2N6028. This specific type was chosen because it has a very small offset voltage, valley current, and peak current. The complete electrical specifications for the D13T2 and the 2N6028 can be seen in Appendix A.

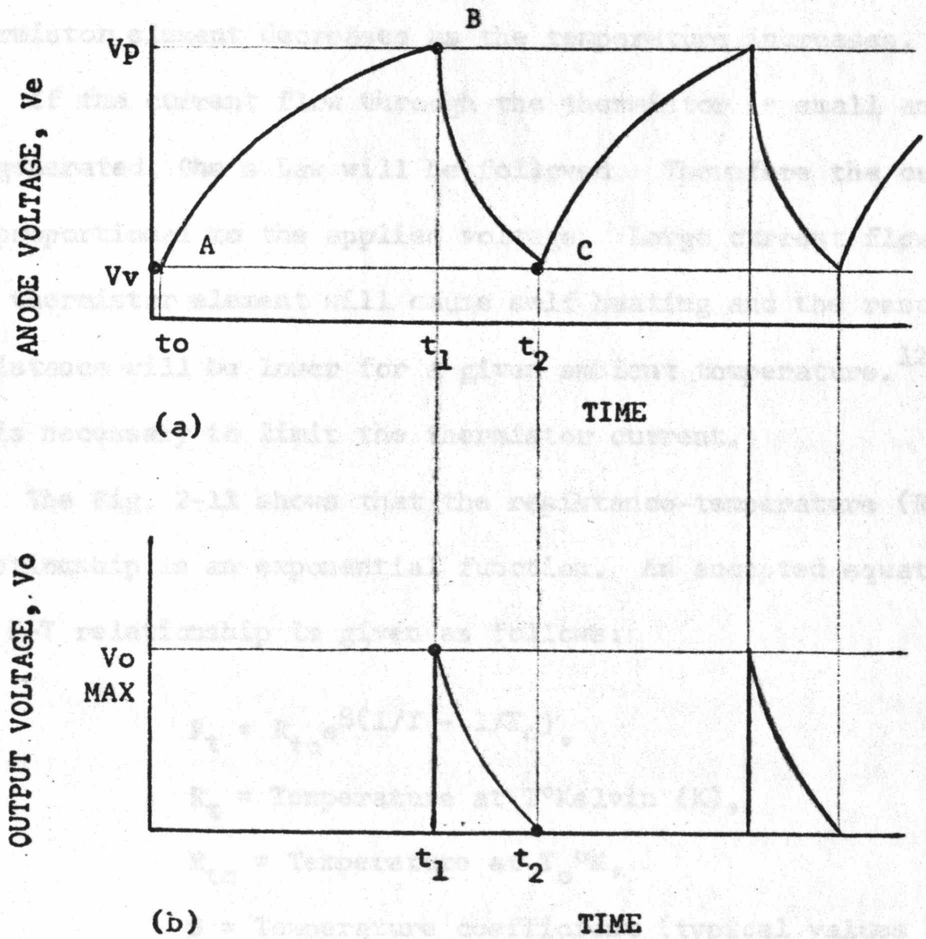


Fig. 2-10. Voltage waveforms of a programmable unijunction transistor relaxation oscillator. (a) Anode voltage waveform (b) Output voltage waveform.

2-3 THE THERMISTOR

As mentioned earlier, a thermistor was used as the temperature sensing element. Thermistors are a resistive semiconductor transducer used to measure temperature.[11] They have a much higher temperature coefficient than pure metals. For example, a typical thermistor has a 3% to 5% change in resistance per degree centigrade (C) change in temperature as compared with .4% for platinum.[12] The coefficient for the thermistor is negative or in other words, the resistance of the thermistor element decreases as the temperature increases.

If the current flow through the thermistor is small and no heat is generated, Ohm's Law will be followed. Therefore the current will be proportional to the applied voltage. Large current flow through the thermistor element will cause self heating and the resulting resistance will be lower for a given ambient temperature.[12] Therefore it is necessary to limit the thermistor current.

The Fig. 2-11 shows that the resistance-temperature (R-T) relationship is an exponential function. An accepted equation for the R-T relationship is given as follows:

$$R_t = R_{to} e^{\beta(1/T - 1/T_o)}, \tag{2-14}$$

R_t = Temperature at T°Kelvin (K),

R_{to} = Temperature at T_o°K,

β = Temperature coefficient (typical values 3000 to 4000),

e = 2.71828.

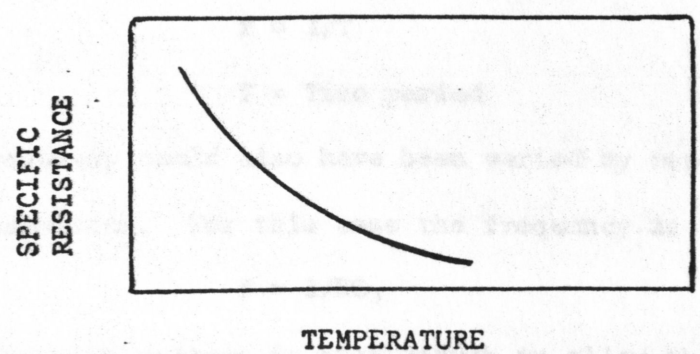


Fig. 2-11. Specific resistance versus temperature of a basic Fenwal Electronics thermistor material.

The thermistor used in the research project was a GA61P8 1 megohm glass Fenwal Electronics thermistor. The glass thermistor consists of a 0.011 inch diameter thermistor bead hermetically sealed in a 0.02 inch X 0.5 inch glass tube. Having the bead sealed in glass isolates it from air, gas, and electrically conductive mediums.[13] See Appendix B for the characteristics of the GA61P8 thermistor.

Getting back to the relaxation oscillator, the thermistor replaces the resistor R1 (Fig. 2-7). Thus a temperature change would result in a change in voltage Vs (Eq. 2-6). According to Eq. (2-10), the peak voltage would be changed directly. As a final result, the change in the temperature of the thermistor would vary the frequency of the oscillator accordingly. The illustration in Fig. 2-12 shows more clearly how changing Vp can result in a change in frequency. Assuming a constant RC time constant, it can be seen that the time period t_1 for peak voltage V_{p1} is much smaller than the time period T_2 for a larger peak voltage V_{p2}. Therefore the frequency (f) for V_{p1} will be greater than the frequency for V_{p2} since frequency is defined as:

$$f = 1/T \qquad (2\text{-}15)$$

T = Time period

The frequency could also have been varied by replacing R (Fig. 2-7) with the thermistor. For this case the frequency is defined

$$f = 1/RC, \qquad (2\text{-}16)$$

providing the peak voltage is high enough to allow the capacitor to charge fully.

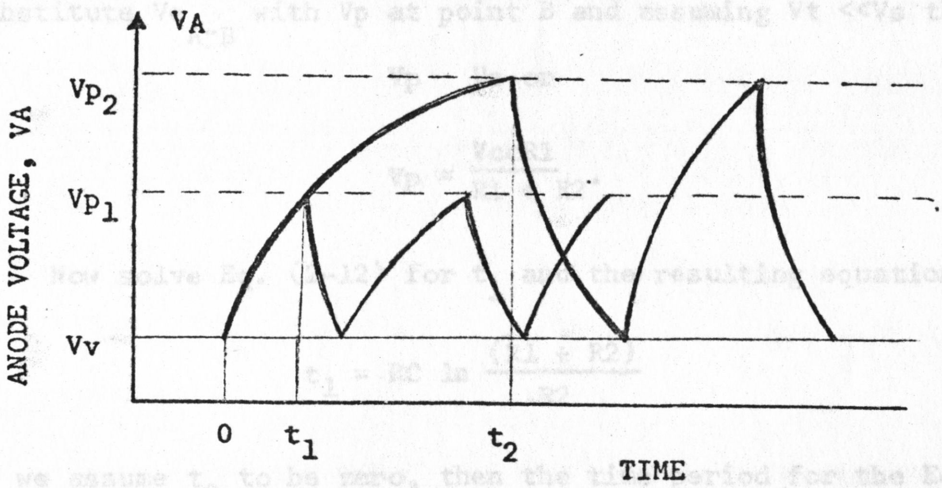


Fig. 2-12. Emitter voltage of a programmable unijunction relaxation oscillator with two values for Vp.

Now with the PUT, thermistor and supply voltage (Sec. 2-4) determined, values for R and C must be calculated next since they have a great influence on the frequency of the oscillator. Since future studies may also include heart rate system, the oscillator frequency should be much greater than the heart beat to avoid any frequency overlaps. A frequency of 8 hertz (Hz) was chosen as minimum frequency. The approximate frequency range of a PUT is between 0.003 Hz and 2.5 Kilohertz (KHz) using capacitor values between 1000 picofarad (pF) and 100 microfarad (uF), respectively.[5] For our purposes a 0.15 uF electrolytic capacitor was chosen. This value would be appropriate for mid-scale frequency use and as discussed earlier, it would help yield maximum output voltage.

With the values for f and C known, we can calculate a value for R with Eq. (2-16). Therefore R is 0.47 megohm.

A more exact equation for f can be derived using Eq. (2-12). Substitute Ve_{A-B} with Vp at point B and assuming Vt <<Vs then we have

Vp Vs or

$$Vp \approx \frac{VccR1}{R1 + R2}. \qquad (2\text{-}17)$$

Now solve Eq. (2-12) for t_1 and the resulting equation will be

$$t_1 = RC \ln \frac{(R1 + R2)}{R2} \qquad (2\text{-}18)$$

If we assume t_o to be zero, then the time period for the Eq. (2-18) would be t_1. According to Eq. (2-15) the frequency of the TSU is $1/t_1$.

Since we replaced resistor R1 with a thermistor, we can substitute R1 with the R_t of Eq. (2-14). The resulting equation can be used in calculating exact frequencies for various temperatures.

The parameters R_{to} and B of Eq. (2-14) were not available for the Fenwal thermistor so Eq. (2-18) was not used for frequency calculation.

A temperature study was conducted to determine the resistance variation of the Fenwal GA61P8 thermistor with respect to temperature. Between the temperature 33 and 44°C the thermistor resistance varied between 0.62 and 0.42 megohm respectively. The above resistance values can be substituted for R1 of Eq. (2-18) to calculate the frequency range for the above temperature range.

At this time a value for R2 is needed so that the $1/t_1$ frequencies can be calculated. R2 is a resistor for the external voltage divider network of the oscillator. Since R1 is the sensitivity element, R2 should be chosen such that R1 would have maximum sensitivity control. Therefore R2 must be approximately equal to R1 or 0.5 megohm.

Actual circuit tests indicated 0.34 megohm would yield maximum output voltage; however, the temperature-frequency response was not linear. The PUT circuit oscillation stops if R2 is less than .34 megohm. In the final design, R2 was 0.57 megohm.

The output voltage for the PUT relaxation oscillator with the above components varied between 0.75 and 1.1 volts with the temperature range between 32 and 45°C, respectively.

The resistor R_o serves as a load resistor and a current limiter so its value is not critical. A value around 12 kilohms or less would suffice for R_o.

The final relaxation oscillator circuit was assembled on the same printed circuit (PC) board as the transmitter as shown in Fig. 2-15. The description of the circuit assembly is included in section 2-4.

2-4 FM TRANSMITTER

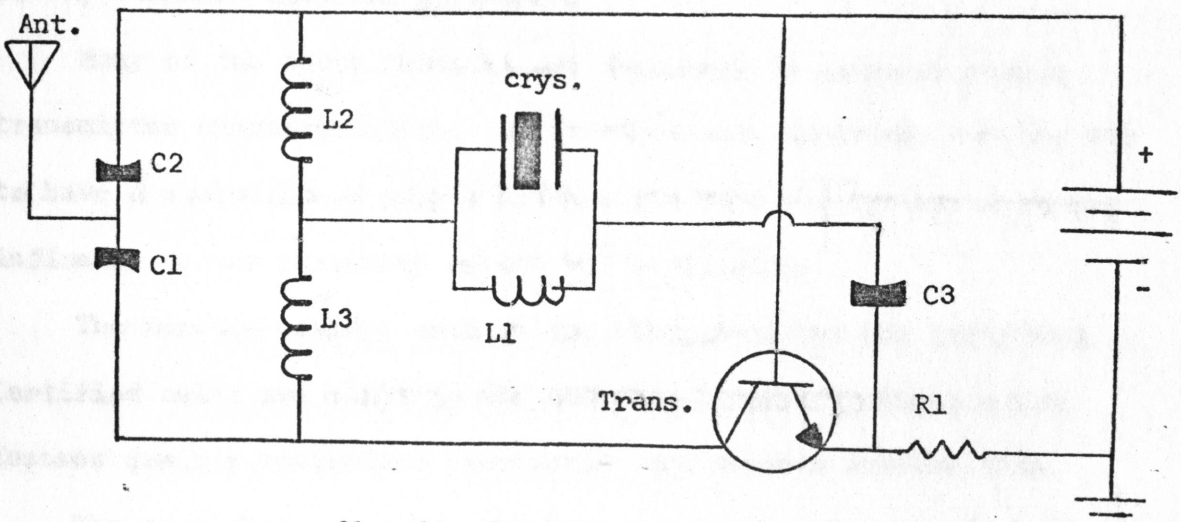
The biotransmitter used in this project is to be frequency modulated. FM was specified because of its high discrimination against noise and interfering signals.[14] The signal-to-noise ratio of FM receivers need only be 2 to 1 as compared to 100 to 1 for that of the AM receiver.[15] The FM circuitry used is very simple since it does not require an additional stage for modulation. With simplicity comes higher reliability and lower power consumption.[16]

The FM frequency range used in short range telemetry is in the area of 100-250 megahertz.[17] These frequencies were chosen because of the advantages of miniaturization.[18] The lower radio frequencies (less than 1 megahertz) give better radio transmission through the body tissue but there are null problems.[19] Such frequencies also require large components and their bandwidth is insufficient. Bandwidth can be described as the frequency between the half-power points of the frequency response curve.[20] When the frequencies are too high (greater than 250 megahertz), packing becomes too critical due to stability problems.[21]

The design of a biotransmitter was not in the scope of this research project. Dr. V. G. Ellerbruch of the Electrical Engineering Department at South Dakota State University designed a 150 megahertz crystal controlled oscillator transmitter for tracking pheasants. Ellerbruch's transmitter met the frequency, size, weight, transmission range, and temperature stability criteria of the biotransmitter desired for this project.

A circuit diagram and component values for the transmitter are shown in Fig. 2-13. Basically, the transmitter is a Hartley oscillator with a crystal used as a tuning element between L2 and L3 to ground.[20,22]

The crystal was built according to specifications dictated by the circuit design. The crystal is a series resonance type. If the crystal is operated at resonance, it will look resistive in the circuit and the crystal impedance will be near zero. The resistance of the M-tron crystal is 50 ohms and the shunt capacitance is 6 pF at resonance. The inductor L1 is shunted across the crystal to reduce the capacitive effect of the crystal to near zero.



C1 = 10 pf L1 = 0.12 μh
C2 = 12 pf L2 = 0.068 μh
C3 = 180 pf L3 = 0.047 μh
Crystal = M-tron quartz 150.065 megahertz
Antenna = 6.5 cm of piano wire
Transistor = Motorola MPS 918 NPN
Battery = 1.38 volts certified Mallory

Fig. 2-13. Ellerbruch's transmitter.

The inductors L1, L2, and L3 are precision built miniature inductors. The capacitors C1, C2, and C3 are mica capacitors which meet rigid moisture requirements.

The current drain of the transmitter is 46.5 microamperes (μA). The current drain of the transmitter is controlled by the gain parameter (h_{FE}) of the transistor. If a similar transistor with a higher h_{FE} were available, the current drain could be reduced and therefore the battery life could be increased. Appendix C contains the maximum and electrical characteristics of the MPS 918.[24]

The power supply for the transmitter is a mercuric oxide battery. The characteristic features of the mercury cell are; flat discharge characteristics, relatively constant ampere-hour capacity, excellent shelf life, good high temperature characteristics, and low and relatively constant internal impedance.[23]

Many of the above features are necessary to maintain stable transmitter characteristics. It is especially important for the TSU to have a stable power supply because the bias voltage has a strong influence on the frequency of the PUT oscillator.

The mercury battery used in the biotransmitter was certified. Certified cells are built to the customers' specifications which insures quality controlled fabrication and minimum storage time.

The certified mercury cells have a capacity of approximately 1000 milliampere-hours. The current drain of the TSU and the biotransmitter were 6 and 46.5 microamperes, respectively. The life of the mercury cell is approximately 3 months.

A discussion of the biotransmitter tuning, temperature stability, and circuit assembly will be included in section 2-5.

2-5 TEMPERATURE SENSITIVE TRANSMITTER UNIT

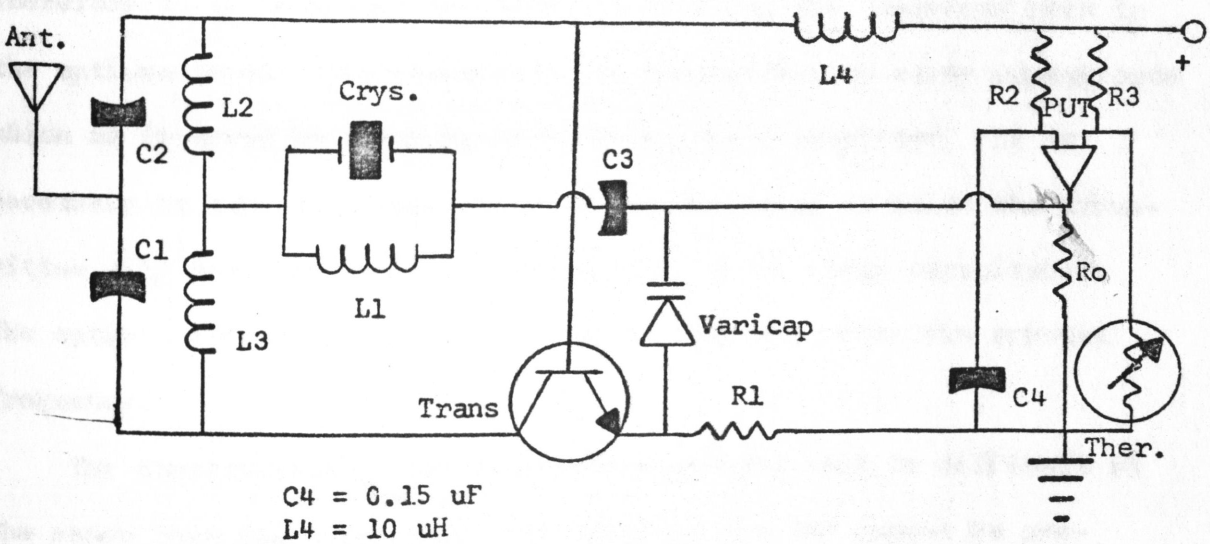
The biotransmitter discussed in section 2-4 can be amplitude modulated or frequency modulated. For reasons discussed earlier, FM was chosen.

To FM the transmitter it is necessary to modulate the carrier frequency more than $\pm$ 5 kilohertz as dictated by the receiver requirements. See chapter III for the radio specifications.

Silicon epicap diodes (varicaps) are used in electronic tuning. The characteristics of the epicap are such that its capacitance varies inversely to applied reverse voltage. A Motorola MV1652 epicap diode with a nominal diode capacitance of 120 pF at a reverse voltage of 4 volts dc and frequency of 1 megahertz was used to FM the biotransmitter.[24]

The varicap was connected between the capacitor C3 and the emitter of the transistor in the biotransmitter. The temperature sensing unit's output was connected to the varicap as shown in Fig. 2-14.

Hinz[25] used a varicap in his biotransmitter. For a detailed discussicn on the design refer to his thesis. Others [1,26,27] have also used varicaps to frequency modulate transmitters.



C4 = 0.15 uF
L4 = 10 uH
R2 = 0.47 megohm
R3 = 0.57 megohm
R4 = 1 megohm Fenwal thermister
R5 = 10 kilohms
Varicap = MV1652
(Other values are given in Fig. 2-13)

Fig. 2-14. Temperature sensing transmitter.

For easy and compact assembly, the temperature sensing transmitter (TST) is mounted on a printed circuit board as shown in Fig. 2-15.

A short whip antenna is used to radiate energy more efficiently. Whip antennas are more efficient than other available antenna types.[28] Piano wire worked well for the antenna because it did not bend or break easily.

After the circuit assembly is completed, the transmitter must be tuned. The transmitter frequency is locked into the crystal frequency, however, capacitor C2 can function as a tuning capacitor by varying the frequency by ± 20 kilohertz. Introducing the varicap into the transmitter circuit causes the frequency to shift about 20 kilohertz. Therefore it is necessary to adjust C2 to bring the frequency back to the optimum level. The transmitter is susceptible to stray capacitance which is incurred by touching or handling the transmitter. It is necessary to tune the transmitter to the frequency at which the transmitter will have the least susceptibility to the stray capacitance. The optimum frequency is usually 7 or 8 kilohertz below the crystal frequency.

The construction of the temperature sensing unit is difficult in the sense that the exact pulse frequency of the TSU cannot be predicted. The component values all vary due to poor tolerances (± 10%), and thus it is impossible to obtain the designed pulse frequency exactly. Normally the pulse frequency of the constructed TSU can be calculated to within ± 2 hertz of the desired frequency. Therefore it is essential to calibrate each TSU after it is constructed to obtain

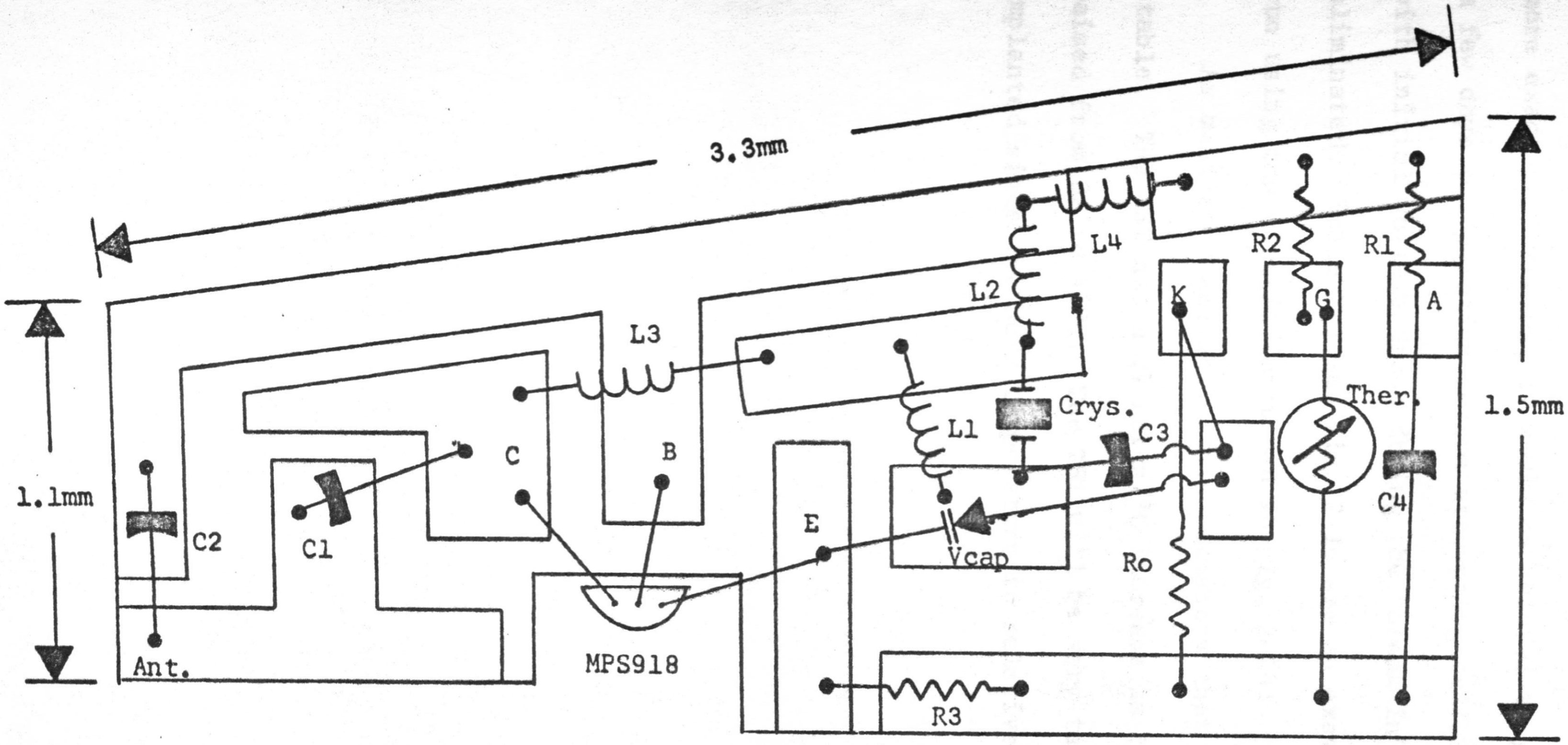


Fig. 2-15. Printed circuit board for the temperature sensing transmitter.

an exact temperature versus pulse frequency characteristic. For a more exact calibration, it would be helpful to give the new battery a few days of operation before attempting calibration. Any problems with initial voltage drops during the "break-in" period would be eliminated. Table 2-1 and Fig. 2-16 are an example of a calibration run using the circuit described in Fig. 2-14.

As mentioned earlier, it is necessary that the TST be temperature stable. The TST had a 37 hertz/°C increase as the temperature was raised from 32 to 45°C. The TST will be very temperature stable when implanted since the body temperature is relatively stable.

TABLE 2-1

Temperature-Frequency Response of Temperature Sensing Transmitter

Temperature (°C) *	Frequency (hertz)
32.2	10.2
32.8	10.64
32.9	11.76
33.9	12.27
35.2	13.70
35.85	14.60
36.5	15.27
37.0	16.39
37.5	16.95
38.3	17.85
38.6	18.52
40.0	20.83

* Transmitter submerged in a water bath

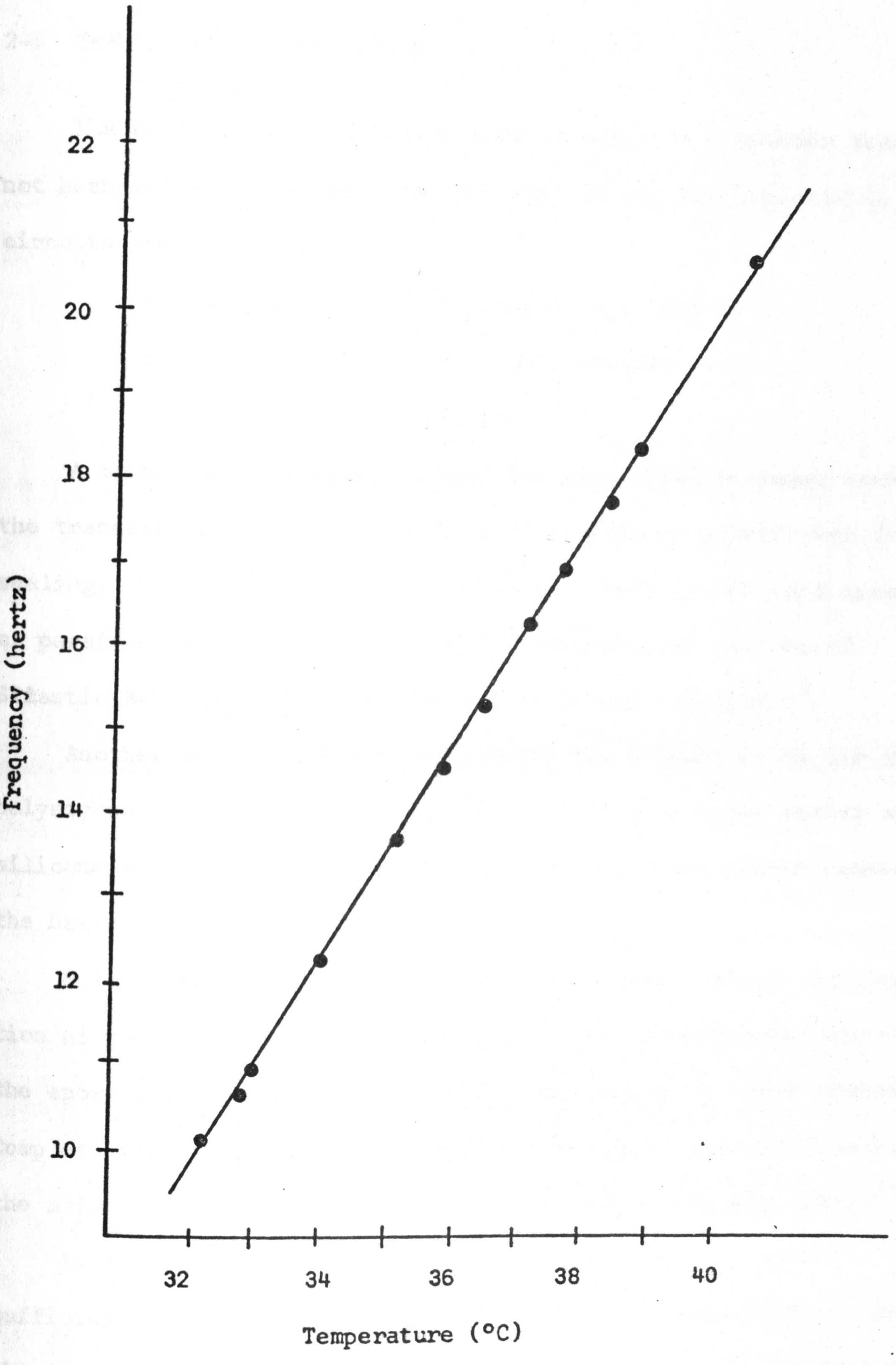


Fig. 2-16. Temperature-frequency response of the temperature sensing transmitter.

2-6 TRANSMITTER ENCAPSULATION

The encapsulation of implantable circuits is a problem that has not been solved. The criteria that must be met for implantable circuits are:

1. high impermeability to saline solutions
2. non-toxic for tissue compatability
3. electrically insulative

A method of encapsulation that has been tried included covering the transmitter with a thin coating of acrylic or polystyrene for sealing. Then the circuit was embedded in Dolp Co-169 hard epoxy or paraffin wax to prevent body fluid penetration. A coat of Silastic 383 was used to prevent tissue incompatability.[17]

Another method consisted of placing the circuit in an air tight polystyrene plastic box. The box was coated with epoxy sealer and silicone rubber. The lactic acid from the silicone rubber caused the battery to corrode.[29]

Potting circuits in epoxy resin only slightly delays the penetration of the saline solutions. A layer of wax 1 millimeter thick over the epoxy can increase the life of the circuit up to three months. Complete encapsulation with silicone rubber approximately doubles the moisture (saline) penetration time of the underlying epoxy.[30]

In general, room temperature curing epoxies do not exhibit sufficient impermeability to moisture. The best results have been obtained with a 50:50 mixture of paraffin and bees wax, covered with a medical grade of silicone rubber for maximum tissue compatibility.[18]

The TST of this project was first encapsulated in a Nu-Weld acrylic denture repair cement. First the circuit was taped with masking tape to prevent the cement contact with the electronic components. This type of encapsulation introduced stray capacitance to the transmitter.

On the second attempt, wax was used to replace the protective masking tape. Whenever wax came in contact with C1, the transmitter would stop transmitting. To prevent wax contact with C1, masking tape was used to cover the capacitor. A coat of Nu-Weld was put over the wax to provide a hard shell. When the transmitter temperature went above 39°C, the wax would melt and flow onto C1 causing transmission failure.

The last encapsulation method included the use of a rubber balloon to cover the entire circuit except the antenna. The balloon functioned as a barrier to keep the encapsulation materials off of the electronic components. A piece of teflon tubing was put over the antenna. The entire TST was encapsulated with Valspar Marine Super Iso-Resin boat fiber-glass repair material. Several coats of the fiberglass were applied to the TST until the shell was approximately 1 millimeter thick. The fiberglass compound consists of 60% unsaturated polyester resin and 40% styrene monomer. For higher permeability characteristics, a .5 millimeter coat of paraffin was applied to the TST. A coat of 383 Silastic medical grade elastomer silicone rubber was applied over the wax tissue for compatibility.

The above encapsulation did not introduce drift nor stray capacitance. High frequency transmitters are more susceptible to signal attenuation than low frequency transmitters, say in the hundreds of kilohertz.[31] Therefore it is necessary to note the effect of the encapsulation on the radiation power and frequency drift. The fiberglass encapsulation had an effect on the TST's radiation power and it will be discussed in Chapter V.

CHAPTER III

BIOTELEMETRY FM RECEIVER

An integral item of the biotelemetry system is the receiver. Although it is essential that the biotransmitter have a clean, strong radiated signal, it is just as important to have a receiver with high sensitivity.

The price range for a receiver is between 100 and 2,000 dollars. Due to the expense of purchasing a receiver, it was decided to use one that the Electrical Engineering Department at South Dakota State University had in stock. The receiver chosen was a Motorola transistorized alert monitor FM receiver. The cost of the receiver is approximately $150.00.

3-1 RECEIVER SPECIFICATIONS

The receiver frequency range lies between 150.8 and 174 megahertz. The carrier frequency of the receiver is set by a crystal in the first oscillator stage and the desired value for the crystal can be calculated as follows:

$$F_{o1} = \frac{Fc - 11.7 \text{ megahertz}}{10}$$

where

F_{o1} = 1st oscillator frequency (crystal frequency),

Fc = carrier frequency.

If TST's of different frequencies were to be monitored with one receiver, it would be necessary to switch crystals to correspond to the frequencies of each TST.

The receiver sensitivity is less than 0.5 microvolt for full 20 db quieting. If the squelch is adjusted to threshold, then the receiver sensitivity is less than 0.25 microvolt. The channel spacing is 30 kilohertz. Either 117 volts ac, 60 cycle or a 12 volts, dc, negative ground source may be used to power the receiver. The receiver modulation acceptance is ± 5 kilohertz.

3-2 RECEIVER OPERATION

The squelch switch stages of the receiver activate the audio stage during the time a signal is received and disable the audio stages during the time no signals are present. The inherent noise generated in the receiver is controlled by the incoming signal. The noise controls the squelch stages. The squelch stages control both the audio preamplier stage and the audio driver stage. If no signal is received, the noise present will raise to a high leveland the squelch stages will cut off (squelch) the audio stages.

When a signal is present the noise level will drop to a low level and thus there will be no input signal at the squelch switch stages. This allows the incoming signal to be amplified. Both the audio preamplifier and the audio driver are now unsquelched.

If no signal is present, the squelch control can be rotated clockwise until the background noise is just squelched. This point is known as the threshold and the receiver sensitivity is at a maximum. Therefore the sensitivity of the receiver will be reduced if the squelch control is turned past the threshold.

In Chapter II it was specified that the TST must be modulated more than $\pm$ 5 kilohertz. You will note that the receiver modulation acceptance is $\pm$ 5 kilohertz. Thus, if the TST signal is modulated more than $\pm$ 5 kilohertz, the audio of the receiver will switch between squelch and normal operation at a rate equal to the pulsation frequency of the TSU. In other words, the audio stage gain will be high when the TST is not modulated; however, there will be no audible output because no information is on the carrier.

When the TST is in the modulated mode, the receiver will not receive a signal since the signal was driven out of the modulation acceptance mode. The audio stages will be squelched and just noise will be heard on the speaker. The noise waveform is shown in Fig. 3-1.

With the volume control the amplitude of the speaker waveform can be varied between 0.4 to 4 volts peak-to-peak. The amplitude specification will be dictated by the pulse frequency-to-voltage converter which is discussed in Chapter IV.

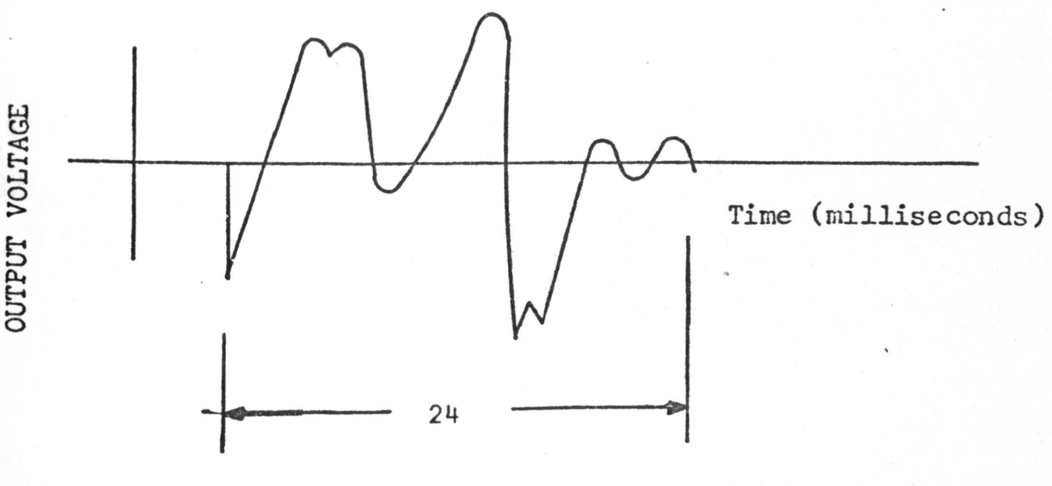


Fig. 3-1. Receiver speaker waveform.

3-3 TRANSMITTER RANGE

The Motorola receiver has an optional external telescopic antenna, however the antenna was not available.

The transmitter was submerged in saline solution to simulate the implant. A glass container with a 10-inch depth and a 10-inch diameter was the bath container. The saline solution temperature was 37°C.

With the radio and the saline solution about six inches off of the ground, the TST range was approximately 50 feet.

When the TST lay horizontal, there was a transmission null at ±90 degrees with respect to the receiver. No nulls were noted when the transmitter was rotated in vertical position.

CHAPTER IV

PULSE FREQUENCY-TO-VOLTAGE CONVERTER

Information concerning variables such as sound, pressure and temperature can be converted into electrical information (temperature) converter and transmitter. The radio receiver receives the transmitted information and converts it to an electrical signal as shown in Fig. 3-1.

Now the electrical information must be converted to readable information. One procedure is to use an analog recorder to record the information. Then the time between pulses can be measured in order to use Eq. (2-15) to calculate the frequency. Next the temperature is read directly from the temperature-frequency graph of the TST.

In order to reduce the labor involved in this conversion, an instrument that would convert the radio output signal into a direct temperature readout is needed.

4-1 CONVERTER SPECIFICATION

The converter must be able to accept and convert non-uniform signals such as those shown in Fig. 3-1. The frequency range of acceptance should be between 5 and 40 hertz. The frequency to temperature response must be linear within that range. Since each TST has a different temperature-frequency slope, the converter must have an adjustable temperature-frequency response slope in addition to an adjustable offset.

The digital temperature readout is to have a range from 32 to 42°C. The converter must have 0.1°C accuracy.

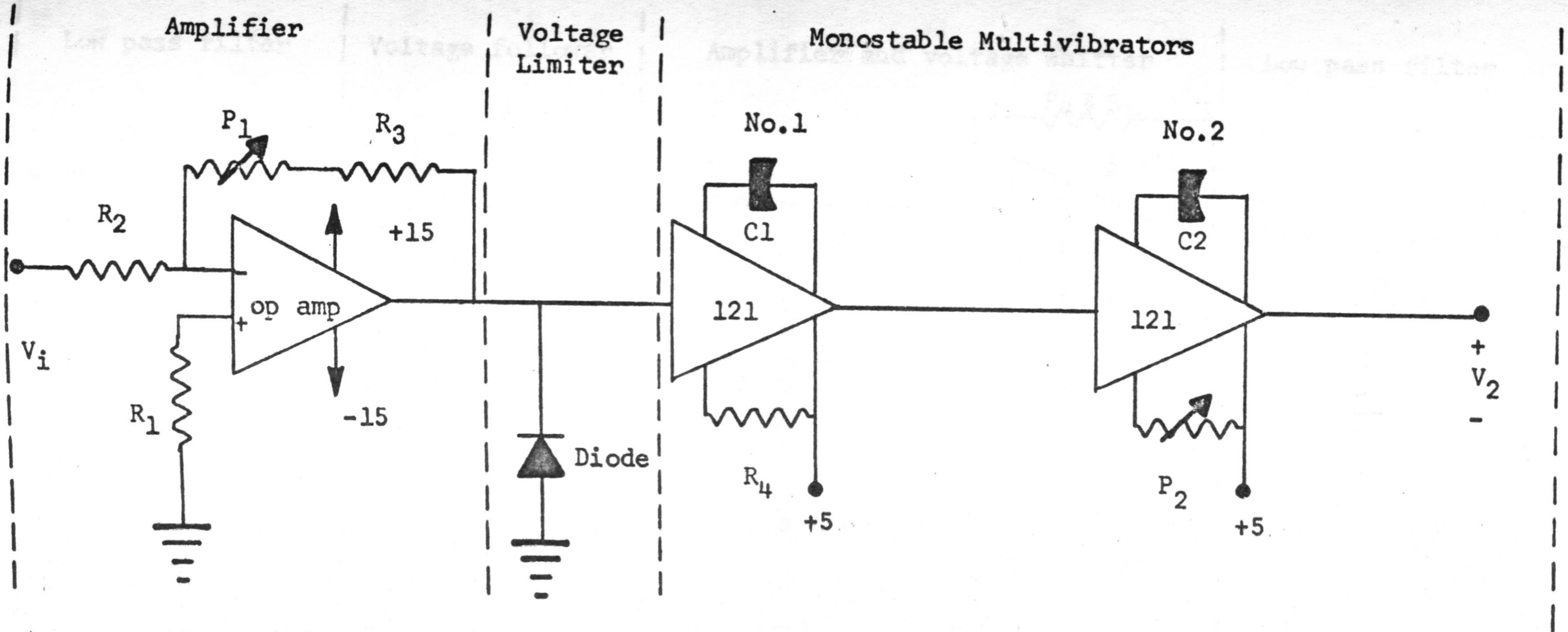
4-2 CONVERTER DESIGN

The converter consists of a wave shaping circuit and a low-pass filter. The wave shaping circuit converts an irregular signal pulse into a pulse of constant height and width. The filter is the key circuit in obtaining a dc voltage proportional to the product of the waveform voltage, waveform frequency, and pulse width. The complete converter system is shown in Fig. 4-1a and Fig. 4-1b.

4-3 INPUT AMPLIFIER

The amplitude of the radio output will not be constant due to transmitter orientation (distance and position). Therefore it is desirable to have the ability to amplify the converter input signal up to 50 times to compensate for nulls and peaks in the received signal. Due to the high input impedance, high gain, low output impedance and versatility of an operational amplifier, it was chosen to amplify the converter's input signal.[32] A dual high performance SN72747 op amp with offset-voltage null capability was chosen. The SN72747 is a pair of SN72741 op amp's constructed on a single monolithic chip. It was chosen for it's availability and low cost. The schematic, maximum rating, and electrical characteristics are given in Appendix D.[33]

The input amplifier is an op amp in the inverting type feedback as shown in Fig. 4-2. In this case the instantaneous output voltage is opposite in polarity to the instantaneous input signal.[34]



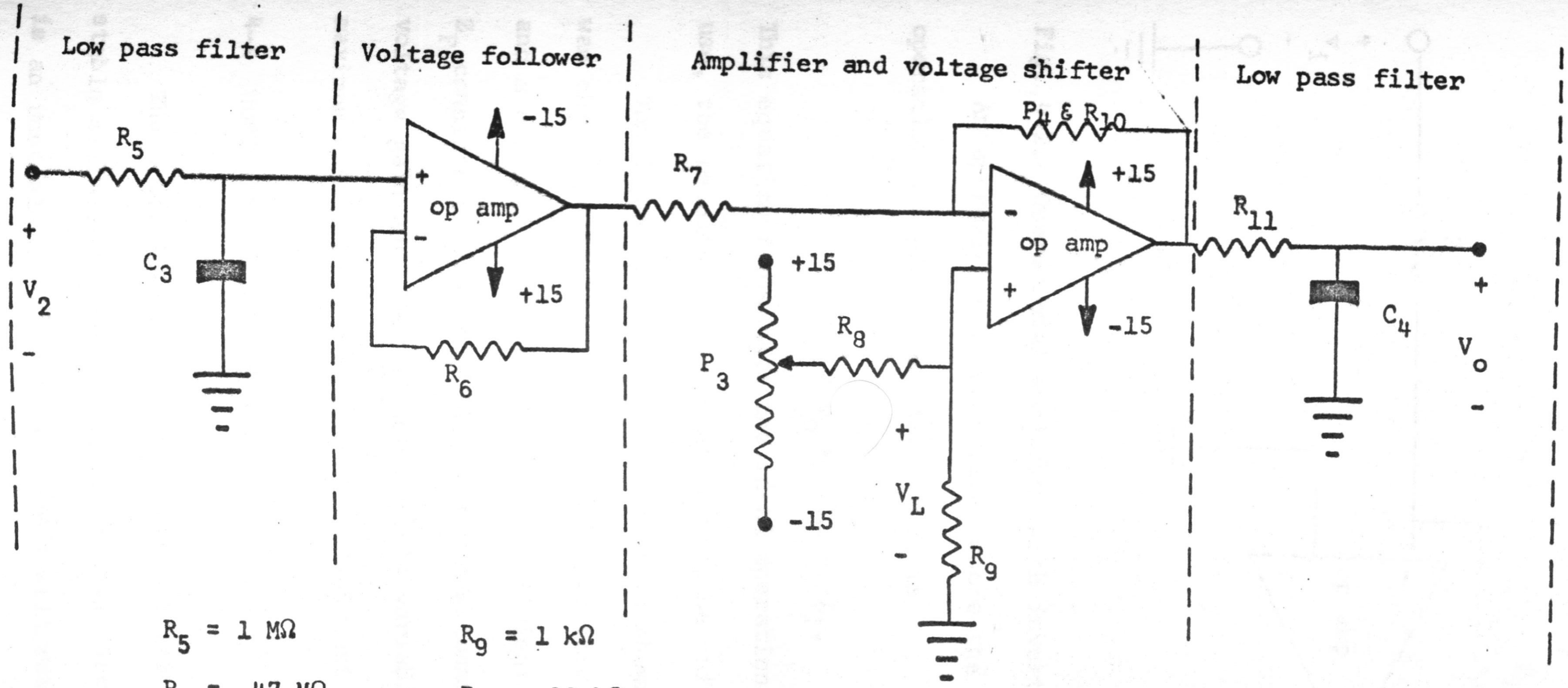
R_1 = 1 kΩ P_1 = 100 kΩ C_1 = .47 uf

R_2 = 10 kΩ P_2 = 10 kΩ C_2 = 4.7 uf

R_3 = 100 kΩ

R_4 = 50 kΩ

Fig. 4-1a. Preamplifier and monostable vibrator waveshaper.



$R_5 = 1\ M\Omega$

$R_6 = .47\ M\Omega$

$R_7 = 10\ k\Omega$

$R_8 = 10\ k\Omega$

$R_9 = 1\ k\Omega$

$R_{10} = 10\ k\Omega$

$R_{11} = 2\ M\Omega$

$C_3 = .48$ uf

$C_4 = .15$ uf

$P_3 = 1\ M\Omega$ Pot.

$P_4 = 100\ k\Omega$ Pot.

Fig. 4-1b. Low pass filter with level shifter.

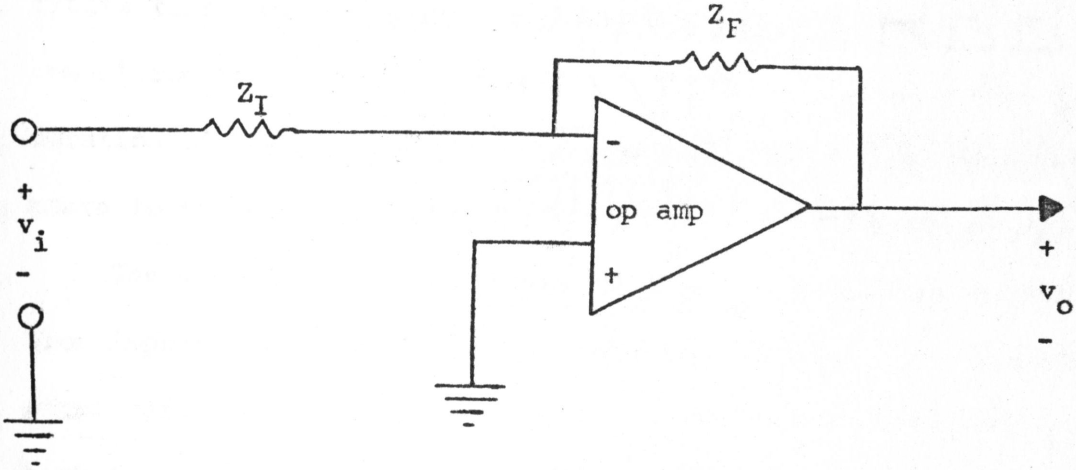


Fig. 4-2. Operational amplifier with inverting feedback.

An approximate forward gain, where the forward gain of the operational amplifier is maximized, is

$$v_o \simeq - Z_F\, v_i/Z_I. \tag{4-1}$$

This equation represents the ideal operational amplifier and for our use, the results of the equation will be sufficiently accurate.

The input operational amplifier is shown in Fig. 4-1a. Z_I(R2) was chosen to be 10 kilohms. A minimum forward voltage gain of 10 and a maximum of 100 would insure sufficient input amplification. Z_F consists of two resistors R3 = 100 kilohms for minimum forward voltage gain and a potentiometer (P1) variable up to 1 megohm for maximum gain. P1 provides a gain adjustment between 10 and 100.

4-4 MONOSTABLE MULTIVIBRATOR

The monostable multivibrator is a binary circuit with only one stable state and one quasi-stable state. The quasi-stable state is an unstable state and the circuit will remain in that state for a finite time duration only. An external signal is required to induce transition from the stable state to the quasi-stable state. The time duration of the quasi-state is adjustable. The return from the quasi-state to the stable state does not require an external signal.

The monostable multivibrator generates rectangular waveforms from impulse or nonuniform short duration signals. The circuit is often referred to as a "one shot" since it returns from the quasi-state back to the stable state after time T.[35]

A common and versatile monostable multibrator is the SN74121. The logic for the SN74121 consists of a negative NOR, a positive AND gate, and a trigger flip-flop as shown in Fig. 4-3.

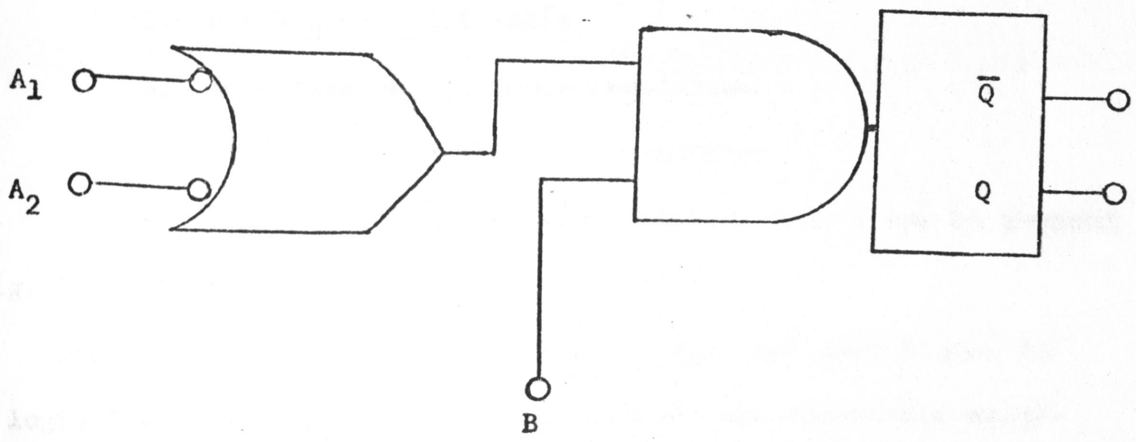


Fig. 4-3. Logic of the SN74121 monostable multivibrator.

The logic truth table for the one shot output of the system is shown in Fig. 4-4.

TRUTH TABLE

t_n Input			t_{n+1} Input			Output
A1	A2	B	A1	A2	B	
0	X	0	0	X	1	one shot
X	0	0	X	0	1	one shot
1	1	1	X	0	1	one shot
1	1	1	0	X	1	one shot

Where:

1. $1 = V_{in}(1) \geq 2$ Volts
2. $0 = V_{in}(0) \leq 0.8$ Volts
3. tn = time before input transition
4. tn+1 = time after input transition
5. X indicates that either a logical 0 or 1 may be present

Fig. 4-4. Logic truth table for the SN74121.

The B input is a positive Schmitt-trigger and when B goes to a logic 1 with either A1 or A2 at a logic 0, the monostable multivibrator will trigger the one shot. A1 and A2 were grounded to maintain a logic 0 at all times so an input voltage of B $\geq$ 2 volts would trigger the one shot.

A zener diode was placed after the input operation amplifier as shown in Fig. 4-1 to limit the input voltage of the monostable multivibrator to 5 volts.

The radio output signal will consist of both a positive and a negative portion which will be amplified as such by the input op amp. The positive section of the signal will always be a logic 1 for B if it is greater than or equal to 2 volts. The negative portion will be a logic 0 for B. Since the radio signal may have more than one positive peak within a period, care must be taken to prevent retriggering of the monostable multivibrator within that same period.

The SN74121 has a feature which enables the variation of the pulse duration T by changing the external timing components. The pulse width can be calculated by the equation

$$T = C_T R_T \log_e 2. \qquad (4-2)$$

To prevent the retriggering of the monostable vibrator, the pulse width must be slightly wider than the positive and negative peaks of the input signal. A pulse duration of approximately 16 milliseconds would be sufficient to prevent such regeneration. Jitter-free operation of the monostable multivibrator may be achieved by staying within the following limits:

$$10 \text{ uf} \geq C_T \geq 10 \text{ pf}$$

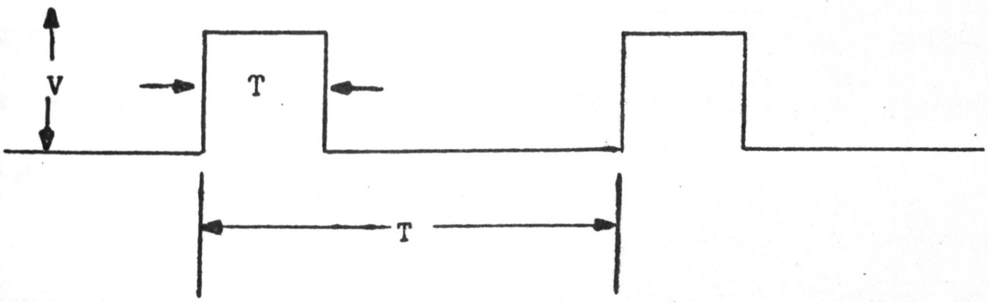
$$40 \text{ k}\Omega \geq R_T \geq 2 \text{ k}$$

With $T \approx 16$ milliseconds and $C_T = 0.47$ uf, calculation of R_T with Eq. (4-2) yields $R_T \approx 100$ kΩ. The true output of the monostable multivibrator is typically + 3.3 volts.

Note in Fig. 4-1 that there are two monostable multivibrators in the converter system. Number 1 is discussed above. Number 2 is added to shape pulse durations less than 16 milliseconds. Since the pulse frequency of the TST may go as high as 40 hertz, the pulse duration of the monostable multivibrator must be limited to 22 milliseconds to prevent pulse overlap. Basically both monostables have the same type of connections except number 2 has C_T = 4.7 uf and R_T = 10 kΩ potentiometer. The pulse duration for these components is adjustable and it ranges from 6.5 to 34.2 milliseconds. The pulse width of number 2 is 20 milliseconds. See Appendix E for monostable SN78121 specifications and characteristics.

4-5 LOW-PASS FILTERS

The retangular signal of the monostable multivibrator will have an average or a dc value. An equation for the dc value may be derived using Fig. 4-5.



T - pulse duration   V - pulse amplitude
T - period

Fig. 4-5. Monostable multivibrator output signal.

From the Fourier series we have

$$a_o = V_{dc} = \frac{1}{T}\int_0^T f(t)dt$$

where

$$v(t) = \begin{cases} V, & 0 < t < T \\ 0, & T < t < T, \end{cases}$$

$$V_{dc} = \frac{1}{T}\int_0^T Vdt + \frac{1}{T}\int_T^{T-t} 0dt$$

and from Eq. (2-15) a substitution for $\frac{1}{T}$ so

$$V_{dc} = VTf.^{36} \qquad (4\text{-}3)$$

The V and the T of the above equation are constants determined by the monostable multivibrator. The frequency f varies according to the temperature-frequency response of the TST being monitored. Therefore the V_{dc} will be directly related to the temperature of the TST. In order to design a converter with direct temperature read-out, it is necessary to determine the V_{dc} from the monostable multivibrator signal.

A low-pass filter as the one shown in Fig. 4-1b is the most direct method that can be used to determine the V_{dc} value of the signal. If the RC time constant of the filter is much greater than the period of the lowest frequency pulse train; or

$$2\frac{1}{f_1} \leq RC \qquad (4\text{-}4)$$

then the low-pass filter output will be proportioned to V_{dc} of the rectangular signal.[10]

The minimum frequency of the TST is 10 hertz and therefore $F_1 = 10$ hertz. From Eq. (4-4) the RC time constant is approximately 50 milliseconds. With R5 = 1 megohm and C3 = 0.48 uf, the RC time constant would be 480 milliseconds and well within criteria limits for effective filtering.

The maximum V_{dc} deviation may be calculated with use of Eq. (4-3) for the TST discussed in Chapter II. The minimum and maximum frequencies of the TST are 10.2 and 20.83 hertz, respectively as shown in table 2-1. V and T are equal to 3.3 volts and 20 milliseconds, respectively. Therefore the V_{dc} deviation is equal to 0.7 volts.

4-6 VOLTAGE FOLLOWER

Isolation of the filter is accomplished by using a voltage follower as shown in Fig. 4-1b. It has the characteristic of high input impedance which is in the order of one megohm.[37]

4-7 V_{dc} AMPLIFIER AND VOLTAGE SHIFTER

The V_{dc} deviation is approximately 0.7 volt/ °C change. Another operational amplifier in the inverting mode amplifies the V_{dc} deviation. Again a potentiometer (P4) is used in the feedback path to adjust gain.

It was mentioned earlier that the voltage output of the converter and the TST must have the same temperature-frequency slope. The potentiometer P4 provides the capability of changing the slope or varying Θ as shown in Fig. 4-6.

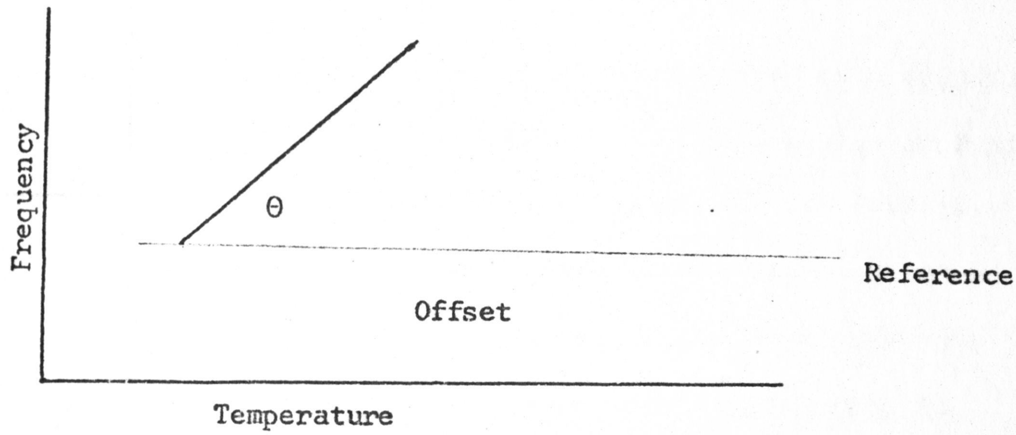


Fig. 4-6. Temperature-frequency characteristic of the converter.

The temperature frequency curve is not only a function of Θ but it is also a function of offset as shown in Fig. 4-6. A voltage shifter (Fig. 4-7) incorporated with the operational amplifier will shift the curve and invert it according to the equation:

$$v_o \approx \frac{-R_F}{R_I}(v_i - V_L), \qquad (4\text{-}5)$$

$$v_o = \frac{-R_F}{R_I} V_i + \frac{R_F}{R_I} V_L .$$

The V_{dc} amplifier and voltage shifter are adjusted such that the output voltage will be one-tenth of the corresponding temperature. For example, if the TST temperature is 37.5°C, then the amplified output will be 3.75 volts. This technique enables an easy adaptation of a digital voltmeter for the temperature readout.

Another low-pass filter is connected to the output of the V_{dc} amplifier for a better output signal rectification.

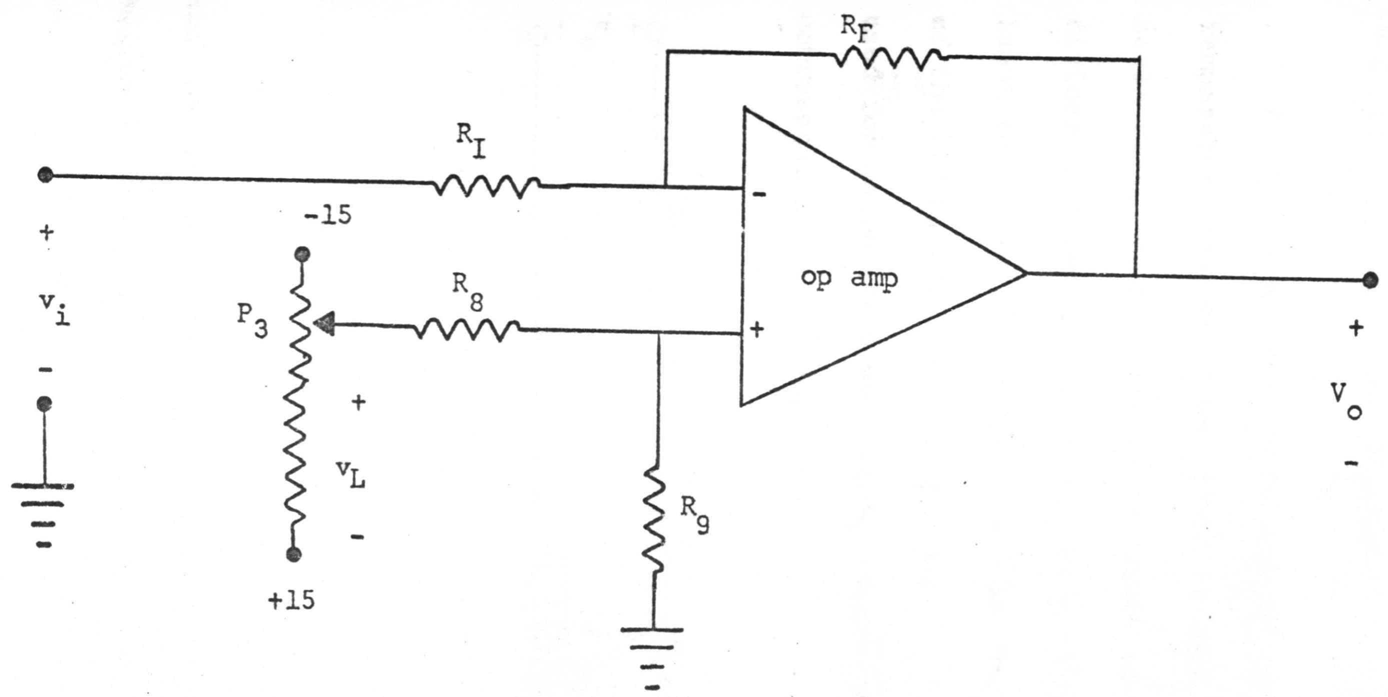
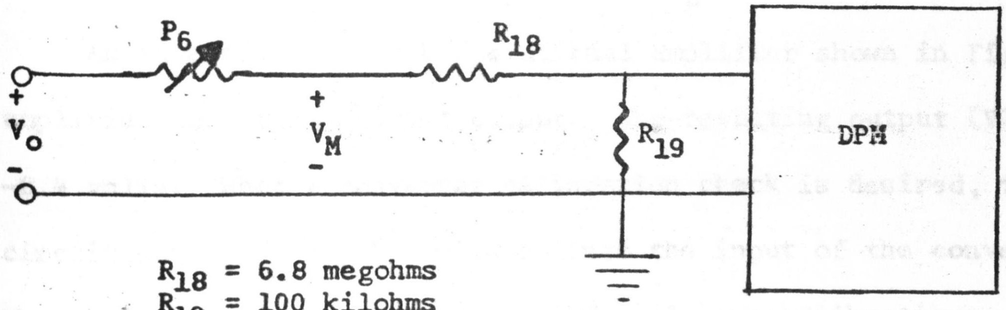


Fig. 4-7. Operational amplifier with a voltage level shifter.

4-8 DIGITAL TEMPERATURE READOUT

An Analog Devices AD2001 digital panel meter serves as a digital temperature readout. The AD2001 is designed for 199.9 millivolt full scale input, however increased ranges may be obtained with a voltage divider as shown in Fig. 4-8. The resistors R18 and R19 boost the input rating to 14.3 volts. Maximum ratings and the specifications of the AD2001 are shown in Appendix F. The potentiometer (P6) is used for calibrating the digital voltmeter so the readout will correspond to the voltage input V_M.



R_{18} = 6.8 megohms
R_{19} = 100 kilohms
P_6 = 137 kolohms

Fig. 4-8. Digital panel meter.

After the slope adjustment discussed in section 4-7 is made, each particular dc voltage output (v_o) will refer to a specific measured temperature. The panel meter has four readout tubes and each tube has an optional decimal point. The decimal point is placed such that the readout is 10 times larger than the input V_M. Therefore, if V_M is 3.75 volts, the actual temperature readout will be 37.5°C. All the voltage-temperature conversions are done in volts to degrees centigrade.

4-9 PUT PULSE CIRCUIT FOR CONVERTER CALIBRATION

For accurate temperature measurement, it is advisable to check the temperature frequency slope of the converter every 24 hours of operation. A PUT pulse circuit as shown in Fig. 4-9 is used as a calibration check. P5 consists of three different resistors which can be switched into the pulse circuit. The resistor-frequency relation is as follows:

P5 = 270 kilohms f_1 = 18.0 hertz

P5 = 437 kilohms f_2 = 12.5 hertz

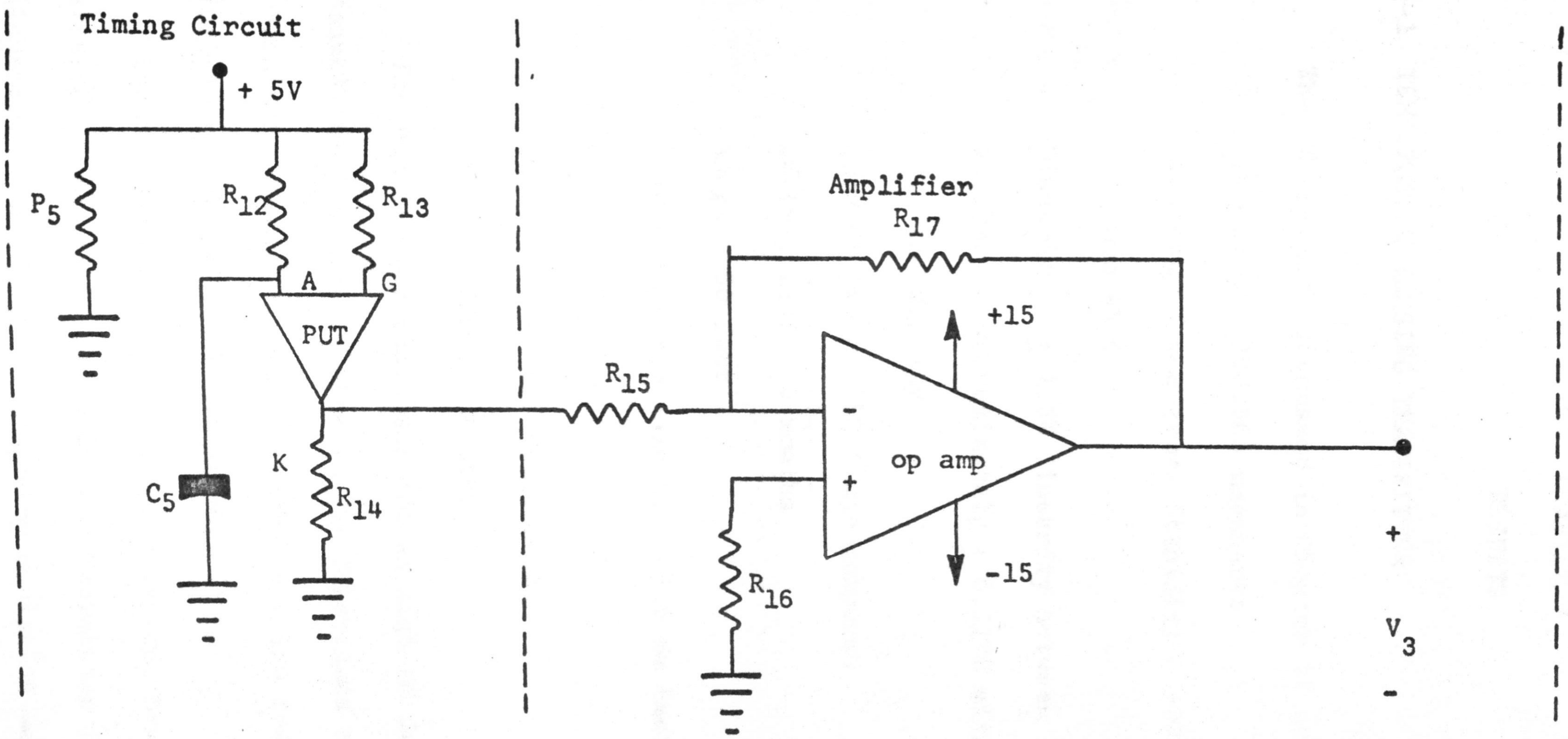
P5 = 800 kilohms f_3 = 8.33 hertz

An inverting feedback operational amplifier shown in Fig. 4-9 amplifies the pulse circuit output. The resulting output (V3) is -8.4 volts. When a converter calibration check is desired, the pulse circuit output V3 may be switched into the input of the converter. The pulse circuit may also be used for slope recalibration.

4-10 CONVERTER POWER SUPPLIES

The SN72747 requires a ± 15 volt supply. A differential supply manufactured by the California Electronics Manufacturing Company is used to power the converter's operational amplifiers.

A 5 volt supply is needed for the monostable multivibrator, the pulse circuit, and the panel meter. A Powertec 5 volt supply with a 3 amperes maximum output was chosen. A ± .2°C jitter occurs when the panel meter, converter, and display share a power supply. For a ± 0.1°C jitter regulation it is helpful to have a separate power supply for the panel meter display. An unregulated supply would suffice.



R_{12} = 1 MΩ
R_{13} = ,57 MΩ
R_{14} = 1 kΩ

R_{15} = 10 kΩ
R_{16} = 1 kΩ
R_{17} = 10 kΩ

C_5 = .15 uf
P_5 = 270 kΩ, 437 kΩ, and 800 kΩ

Fig. 4-9. Timing circuit.

CHAPTER V

RESULTS

5-1 TEMPERATURE SENSING TRANSMITTER

The TST results discussed in Chapter II are tabulated below:

Frequency - 150.065 megahertz

Frequency Temperature Stability - +37 hertz/°C between 32 and 45°C

Linearity - ± 1.3% linearity between 32 and 40°C (Fig. 2-16)

Accuracy - approximately ± 0.11°C with a separate supply for the display

Current Drain - 52.5 microamperes

Battery Life - 3 months

Range - 50 feet

Size - Approximately 6.1 X 1.5 mm (not including antenna)

Weight - 2 ounces

Cost - 35 dollars (materials)

Battery Life - 3 months

The transmitter range was not as high as desired. Before the transmitter was encapsulated in the fiberglass encapsulation, the TST transmission range was much greater than 100 feet. Apparently the fiberglass reduces power radiation.

Due to varying battery characteristics, newly installed TST batteries increase or decrease the transmitter frequency up to 3 kilohertz. A 3 kilohertz frequency deviation makes the transmitter susceptible to stray capacitance. Therefore it is necessary to retune the transmitter after each new battery installation. Since each new battery has different characteristics, TST recalibration will also be required with every battery change.

5-2 PULSE FREQUENCY CONVERTER

The most important quality that the converter must possess is linearity. The results on a test for linearity are shown in TABLE 5-1 and Fig. 5-1.

Some difficulty was encountered with the potentiometers for the level shifter and V_{dc} operational amplifier feedback. Higher precision potentiometers would be helpful.

5-3 TELEMETRY SYSTEM USES AND POSSIBILITIES

The transmitter can be implanted in animals as large as or larger than the average domestic cat. The animal must be restricted to an area that has a radius less than 50 feet.

If more than one TST of different frequencies are to be monitored simultaneously, the crystals of the first oscillator stage of the receiver monitor would have to be switched to correspond to the proper TST frequency. The converter could be modified by installing and calibrating a converter circuit for each TST. A manual switching function could switch the desired crystal and converter network in for each TST. Only one digital meter would be needed for the multi-system.

The pulse-to-frequency converter could be used as a heart rate meter. Since most heart rates are less than 1.8 hertz, it would be necessary to replace the converter's RC low pass filter with an active low pass filter for effective averaging.

The converter is very linear between 10 and 20 hertz or a corresponding temperature range between 32 and 40°C. Occasionally a temperature deviation of 0.1°C is noted which may be due to network drift or calibration inaccuracies.

The TST discussed in Chapter II was calibrated. Then the converter was calibrated according to the TST temperature-frequency response. A telemeter system test was conducted with the TST in a water bath. The results can be seen in TABLE 5-2 and Fig. 5-2. The TST response was not plotted on Fig. 5-2 because its response did not deviate more than $\pm$ 0.1°C from the converter response. The TST was not linear below 33°C or above 42°C.

TABLE 5-1

Converter linearity test

Converter Input (hertz)	Frequency Deviation (Δf)	Temperature Readout (°C)	Temperature Deviation (ΔT)
10		32.9	
	1		.7
11		33.6	
	1		.8
12		34.4	
	1		.5
13		34.9	
	1		.5
14		35.4	
	1		.8
15		36.2	
	1		.8
16		37.0	
	1		.5
17		37.5	
	1		.9
18		38.4	
	1		.5
19		38.9	
	1		.9
20		39.5	

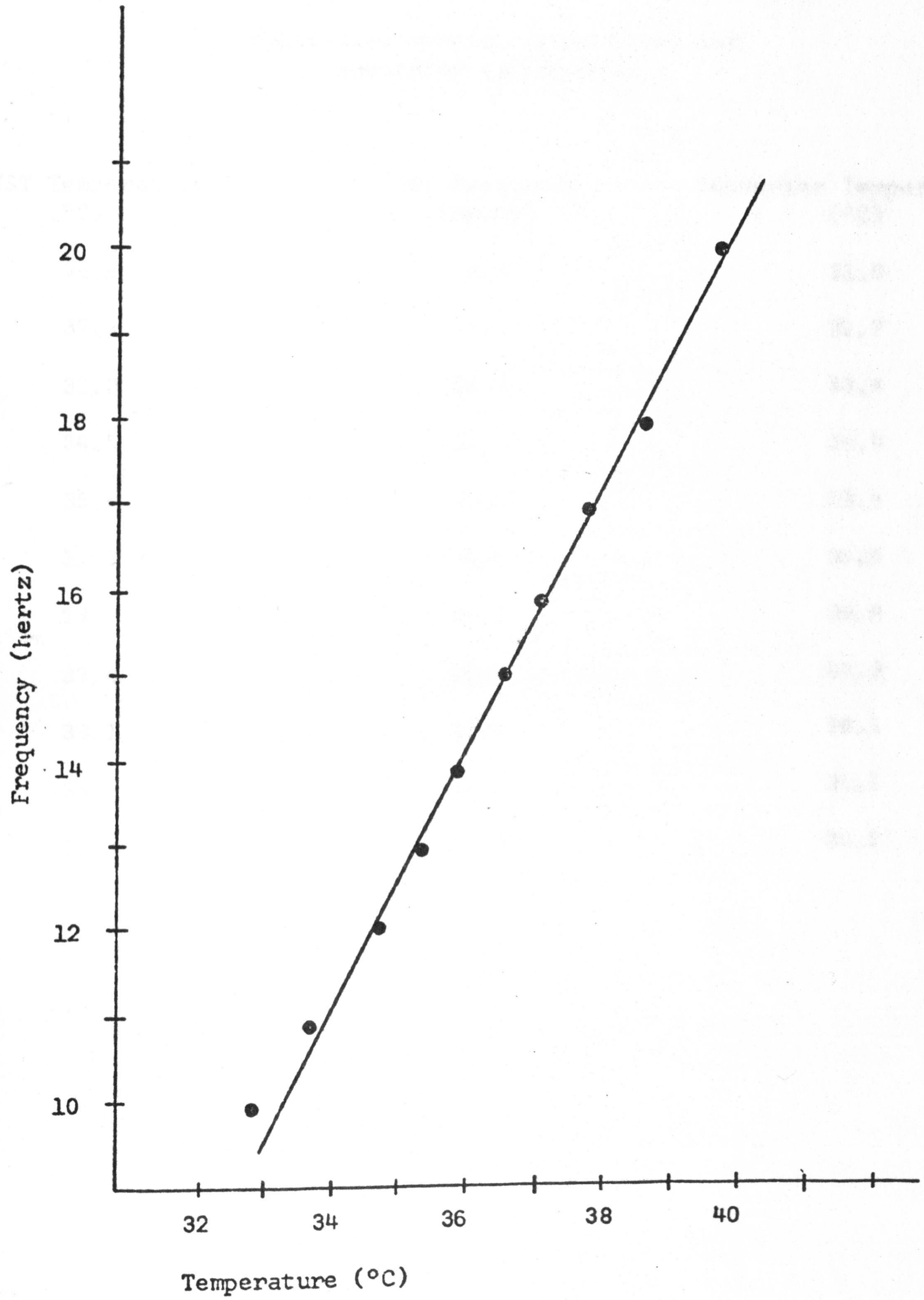


Fig. 5-1. Converter Temperature-frequency Response.

TABLE 5-2

Temperature sensing transmitter and converter calibration

TST Temperature (°C)	TST Frequency (hertz)	Converter Temperature (°C)
30.9	8.7	31.9
32.9	9.8	32.7
33.3	10.9	33.4
34.5	12.5	34.4
35.6	13.9	35.4
36.6	15.6	36.5
37.0	16.1	36.9
37.4	16.7	37.3
38.1	17.9	38.1
38.7	18.5	38.6
39.9	20.8	39.9

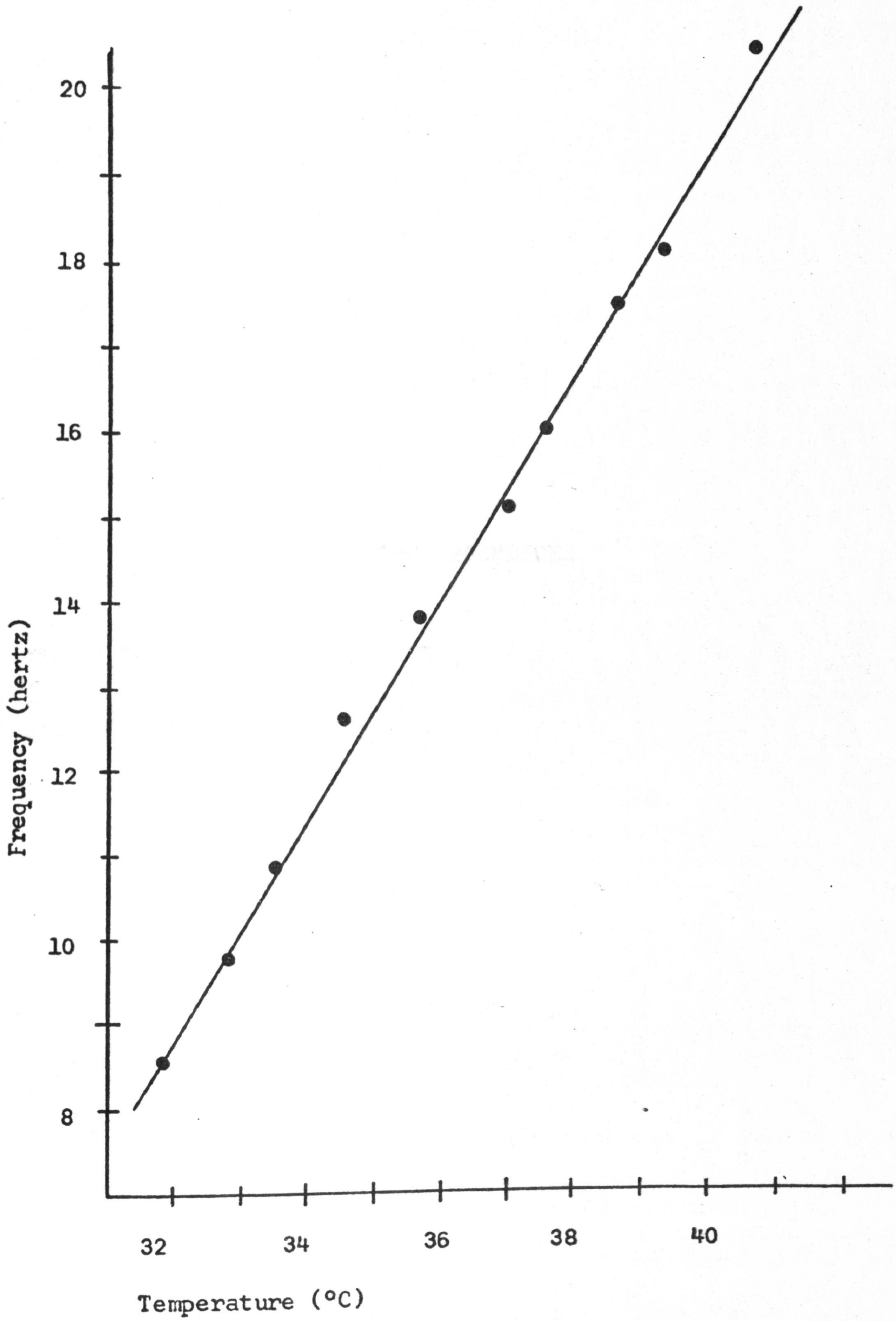


Fig. 5-2. Converter Temperature-frequency Response calibrated to match the TST response.

THE APPENDICES

DEFINITION OF SYMBOLS USED IN THE APPENDICES

V - Volts

mV - Millivolts

A - Ampere

mA - Milliampere

uA - Microampere

nA - Nanoampere

T - Temperature

°C - Degrees centigrade

W - Watt

mW - Milliwatt

Ω - Ohms

Meg - Megohms

k - Kilohms

s - Second

ms - Millisecond

us - Microsecond

ns - Nanosecond

F - Farad

uF - Microfarad

pF - Picofarad

I - Current

APPENDIX A

2N6028 PROGRAMMABLE UNIJUNCTION TRANSISTOR CHARACTERISTICS[7]

ABSOLUTE MAXIMUM RATINGS: (25°C)

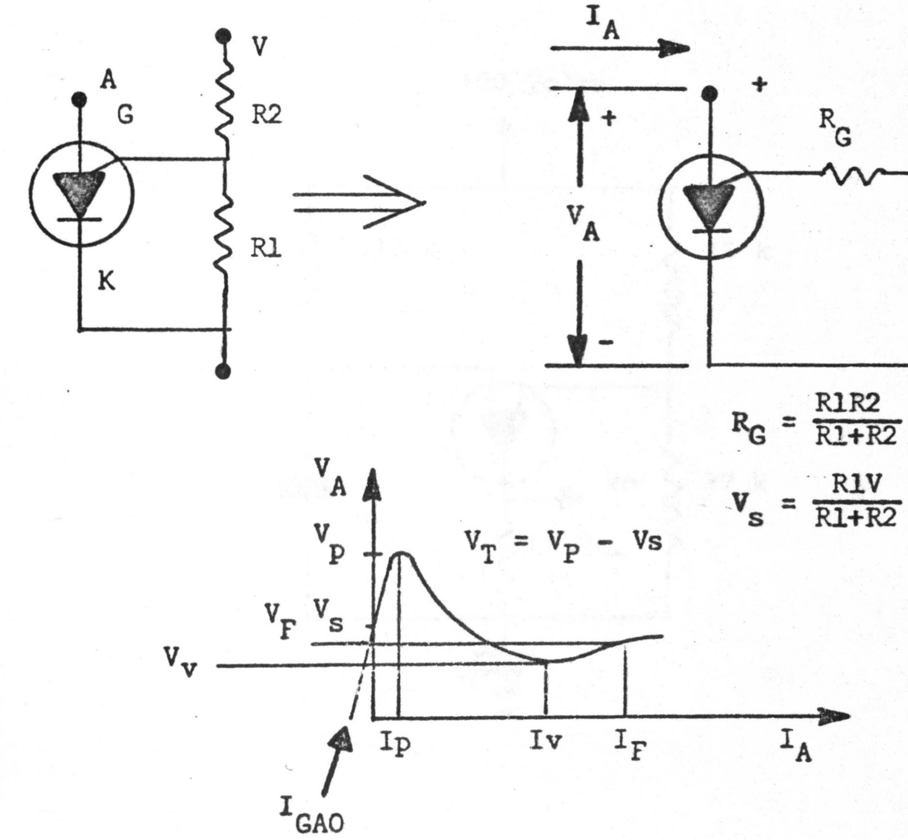
		Units
Gate-Cathode Forward Voltage	+40	V
Gate-Cathode Reverse Voltage	-5	V
Gate-Anode Reverse Voltage	+40	V
Anode-Cathode Voltage	±40	V
DC Anode Current	150	mA
Peak Anode, Recurrent Forward (100 usec pulse width, 1% duty cycle)	1	A
(20 usec pulse width, 1% duty cycle)	2	A
Peak Anode, Non-recurrent Forward (10 usec)	5	A
Gate Current	±20	mA
Total Average Power	300	mW

ELECTRICAL CHARACTERISTICS: (25°C) (unless otherwise specified)

	symbol	Fig.	Min.	Max.	Unit
Peak Current (Vs = 10 Volts)	I_P	A-1			
(R_G = 1 Meg)				.1%	uA
(R_G = 10 k)				1.0	uA
Offset Voltage (Vs = 10 Volts	V_T	A-1			
(R_G = 1 Meg)			.2	.6	V
(R_G = 10 k)			.2	.6	V
Valley Current (Vs = 10 Volts)	I_V	A-1			
(R_G = 1 Meg)				25	uA
(R_G = 10 k)				25	uA
Anode Gate-Anode Leakage Current	I_{GAO}	A-2			
Vs = 40 Volts, T = 25°C)				10	
T = 75°C)				100	nA

Gate to Cathode Leakage Current Vs = 40 Volts, Anode- cat. short	I_{GKS}	A-3		1.5	V
Forward Voltage	V_F	A-3		1.5	V
Pulse Output Voltage	V_o	A-4	6		V

APPENDIX A continued



$$R_G = \frac{R1R2}{R1+R2}$$

$$V_s = \frac{R1V}{R1+R2}$$

Fig. A-1.

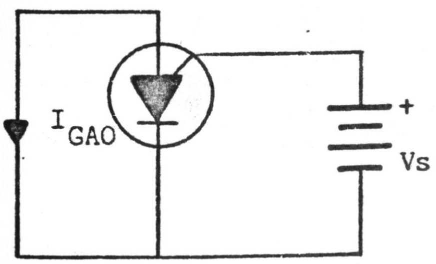


Fig. A-2.

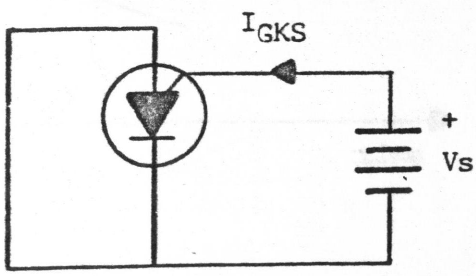
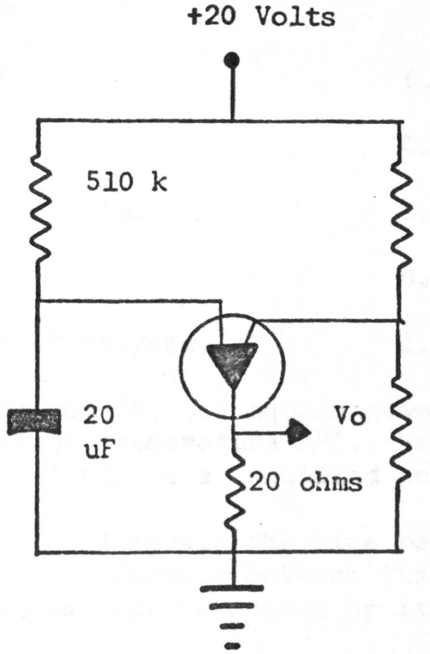


Fig. A-3.

APPENDIX A continued



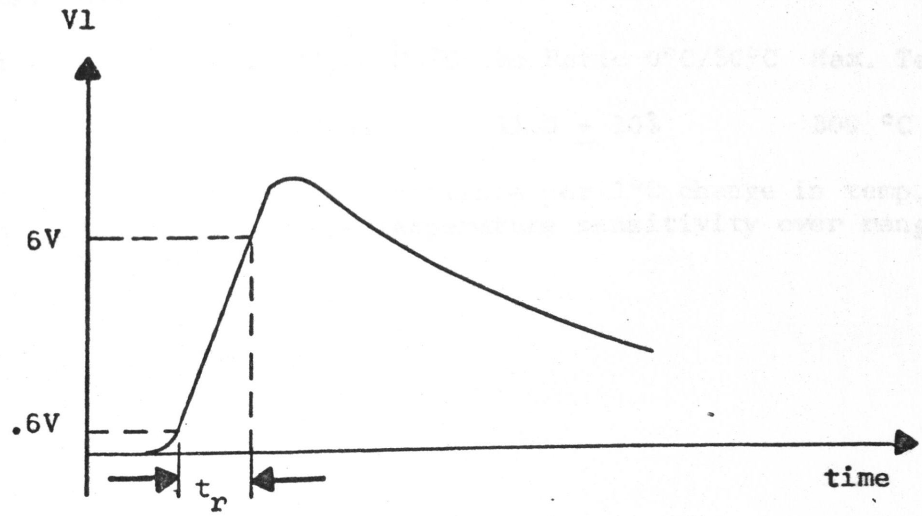


Fig. A-4.

APPENDIX B

FENWAL THERMISTOR CHARACTERISTICS GA61P8[13]

CHARACTERISTICS

CONDITION	D. C. (Nominal)	T. C. (Nominal)
Still Air	0.15 mW/°C	1.4 seconds
Moving Air 700 ft./min.	0.40 mW/°C	0.6 seconds
Still Water	0.75 mW/°C	55 millisec.
Moving Water at 20 ft./sec.	1.25 mW/°C	20 millisec.

Dissipation Constant (D. C.) equals power in milliwatts required to raise thermistor temperature 1°C. Measured with thermistor suspended by its leads in a specified environment.

Time Constant (T. C.) equals the time required by a thermistor to change 63% of the difference between its initial and final temperature. Measured with thermistor suspended by its leads in a specified environment.

TYPICAL UNIT

Ro at 25°C	± % Tol.	Alpha 25°C	Ro Ratio 0°C/50°C	Max. Temp.
1 Meg.	20	-4.9%	11.0 ± 10%	300 °C

Alpha: Percent change in resistance per 1°C change in temp.
Ratio: Measure of average temperature sensitivity over range 0° to 50°C

APPENDIX C

MPS918 CHARACTERISTICS[24]

MAXIMUM RATINGS

RATING	SYMBOL	UNIT
Collector-Base voltage	V_{CB}	Volts
Collector-Emitter voltage	V_{CEO}	Volts
Emitter-Base voltage	V_{EB}	Volts
Total device Dissipation at T_A = 25°C	P_D	mW

ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted)

CHARACTERISTICS	SYMBOL	MIN	MAX	UNIT
Collector Cutoff Current V_{CB} = 15V, I_E = 0	I_{CBO}	-	10	nA
Collector-Base Breakdown Voltage I_C = 1.0 uA, I_E = 0	BV_{CBO}	30	-	Volts
Emitter-Base Breakdown Voltage I_E = 10 uA, I_C = 0	BV_{EBO}	3.0	-	Volts
Collector-Emitter Voltage I_C = 3.0 mA, I_E = 0	BV_{CEO}	15	-	Volts
DC Current Gain V_{CE} = 1V, I_C = 3mA	h_{FE}	20	-	-
Collect-Emitter Sat. Voltage I_C = 10 mA, I_B = 1mA	$V_{CE(SAT)}$	-	0.4	Volts
Base-Emitter Saturation Voltage I_C = 10 mA, I_B = 1mA	$V_{BE(SAT)}$	-	1.0	Volts
Small Signal Current Gain I_C = 4mA, V_{CE} = 10v, f = 100MHz	h_{FE}	6.0	-	-
Amplifier Power Gain I_C = 6mA, f = 200MHz, V_{CB} = 12V	G_{pe}	15	-	dB

APPENDIX D

SN72747 OPERATIONAL AMPLIFIER CHARACTERISTICS [33]

ABSOLUTE MAXIMUM RATINGS OVER OPERATING FREE-AIR TEMP. RANGE

		UNIT
Supply voltage $V_{CC}+$	18	V
Supply voltage $V_{CC}-$	-18	V
Differential input voltage	± 30	V
Input voltage	± 15	V
Voltage between either offset null terminal (N1/N2) and V_{CC}	± 0.5	V
Duration of output short circuit	unlimited	
Continuous total dissipation at free-air temperature	800	mW
Operating temperature range	0 to 70	°C
Storage temperature range	300	°C

ELECTRICAL CHARACTERISTICS AT SPECIFIED FREE-AIR TEMPERATURE
$V_{CC}+ = 15V$, $V_{CC}- = -15V$, 25°C

	PARAMETER	TEST CONDITIONS	TYPICAL	UNIT
V_{10}	Input offset voltage	$R_S \leq 10k$	1	mV
V_{10}	Offset voltage adj. range		± 15	mV
I_{10}	Input offset current		20	nA
I_{IB}	Input bias current		80	nA
V_I	Input voltage range		± 13	V
r_i	Input resistance		2	MΩ
r_o	Output resistance		75	Ω
P_D	Power dissipation	No load	50	mW
I_{CC}	Supply current	No load	1.7	mA

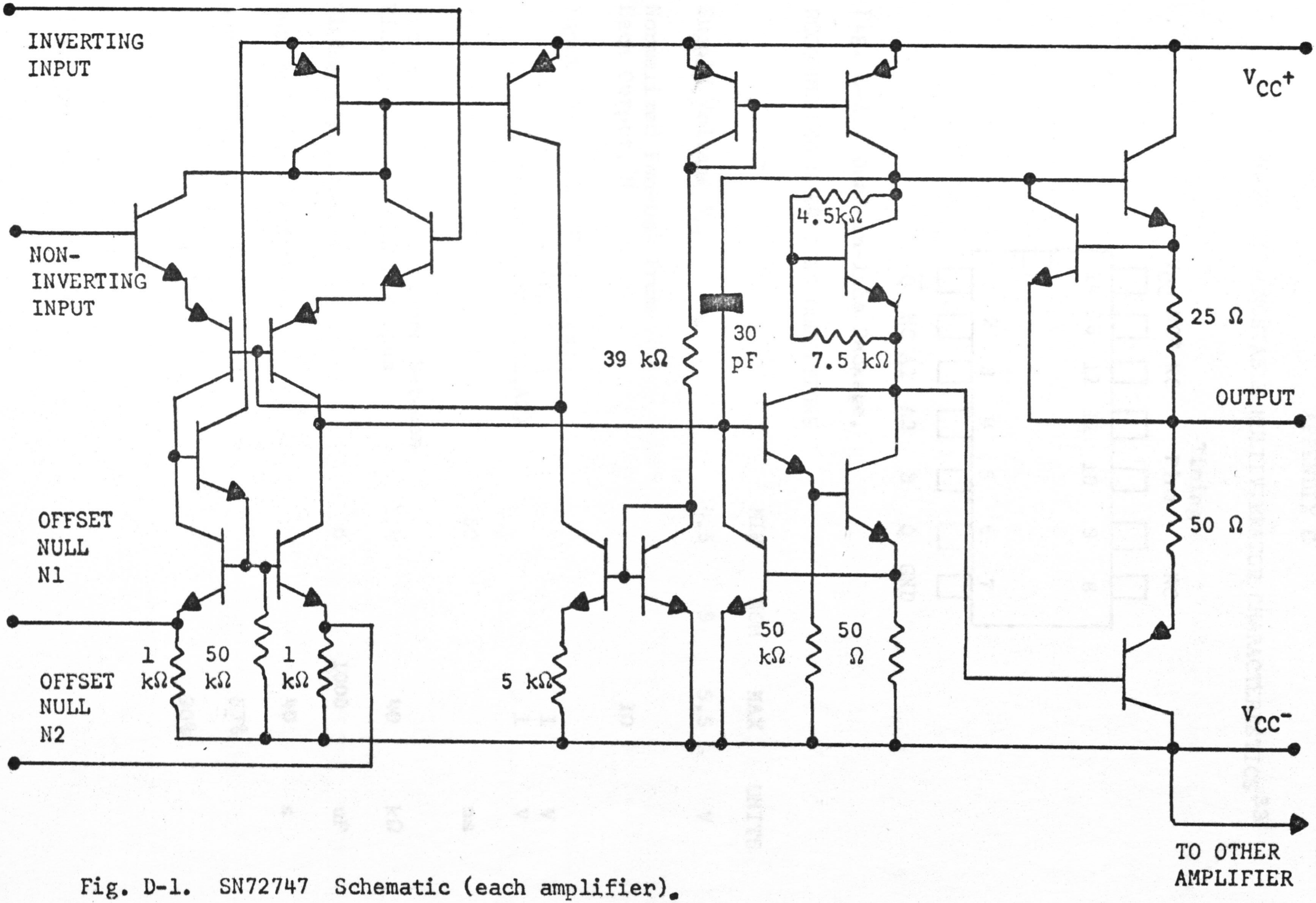


Fig. D-1. SN72747 Schematic (each amplifier).

APPENDIX E

SN72121 MONOSTABLE MULTIVIBRATOR CHARACTERISTICS[33]

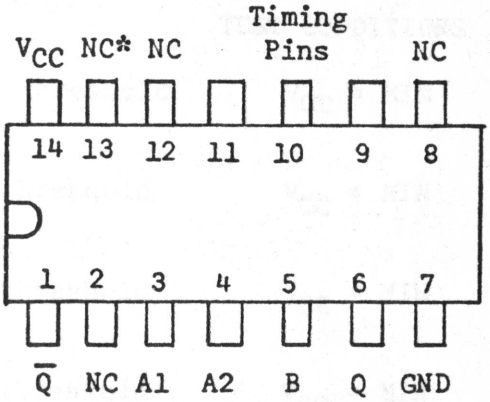


Fig. E-1. Dual-in-line Package.

RECOMMENDED OPERATING CONDITIONS

	MIN	NOM	MAX	UNITS
Supply Voltage V_{CC}	4.5	5	5.5	V
Normalized Fan-Out From Each Output, N			10	
Input Pulse Rise/Fall Time:				
Schmitt Input			1	V
Logic Inputs (A1,A2)			1	V
Input Pulse Width	50			ns
External Timing Resistance Between Pins 11 and 14 with p 9 open	1.4		40	kΩ
Timing Capacitance	0		1000	uF
Output Pulse Width			40	s
Duty Cycle: R_T = 2kΩ			67%	
R_T = 40kΩ			90%	

* No Internal Connection

APPENDIX E CONTINUED

ELECTRICAL CHARACTERISTICS OVER OPERATING FREE-AIR RANGE
USING TEST CIRCUIT OF FIG. E-2.

PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNITS
V_{T+} Positive-going threshold voltage at A input	V_{CC} = MIN		1.4	2	V
V_{T-} Negative-going threshold voltage at A input	V_{CC} = MIN	0.8	1.4		V
V_{T+} Positive-going threshold voltage at B input	V_{CC} = MIN		1.55	2	V
V_{T-} Negative-going threshold voltage at B input	V_{CC} = MIN	0.8	1.35		V
$V_{out(0)}$ Logical 0 output voltage	V_{CC} = MIN, I_{sink} = 16 mA		0.22	0.4	V
$V_{out(1)}$ Logical 1 output voltage	V_{CC} = MIN, I_{load} = -400 uA	2.4	3.3		V

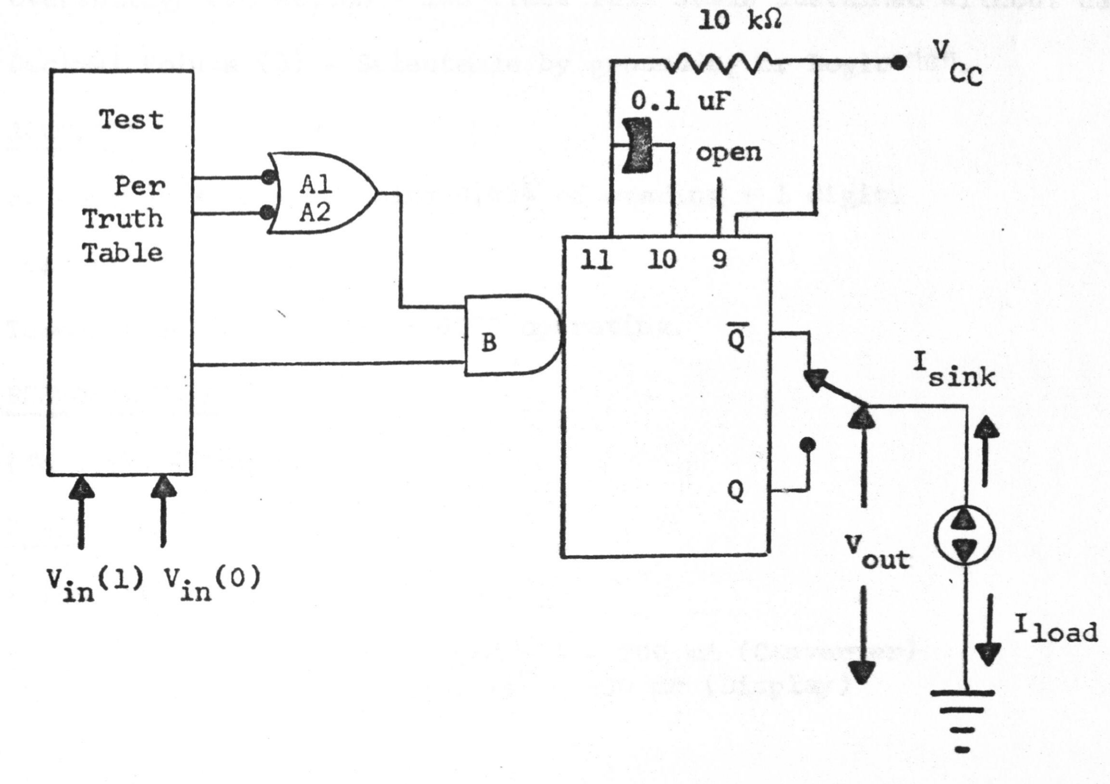


Fig. E-2. d-c test circuit.

APPENDIX F

AD2001 DIGITAL PANEL METER

AD2001 SPECIFICATIONS AT 25°C and 5 volt SUPPLY

DISPLAY OUTPUT

Display consists of 4 RCA Numitrons (7 segment incandescent readout tubes) for data digits plus 100% overrange and ± polarity indication.

Overvoltage-3 data digits read "0"s and the overrange "1" continuously flashes when reading exceeds 200% of full scale.

Decimal Points - selectable at input connector.

INPUT

Full Scale Range - 0 to ± 199.9 millivolts, automatic zero and polarity.

Bias Current - <1 nA.

Impedance - 1000MΩ.

Overvoltage Protection - 100 times full scale sustained without damage.

Decimal Points (3) - Selectable by grounding or Logic "0".

ACCURACY

60 days at 25°C - DC Volts 0.05% of reading ± 1 digit.

Resolution - 0.1 mV.

Temperature Range - 0° to 60°C operating.

RESPONSE TIME

(To 0.5%) - 250 ms

POWER

+5VDC ±5% - 1000 mA, Watts max.

Separate DC inputs - regulated ±5% - 200 mA (Converter)
unregulated - 800 mA (Display)

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